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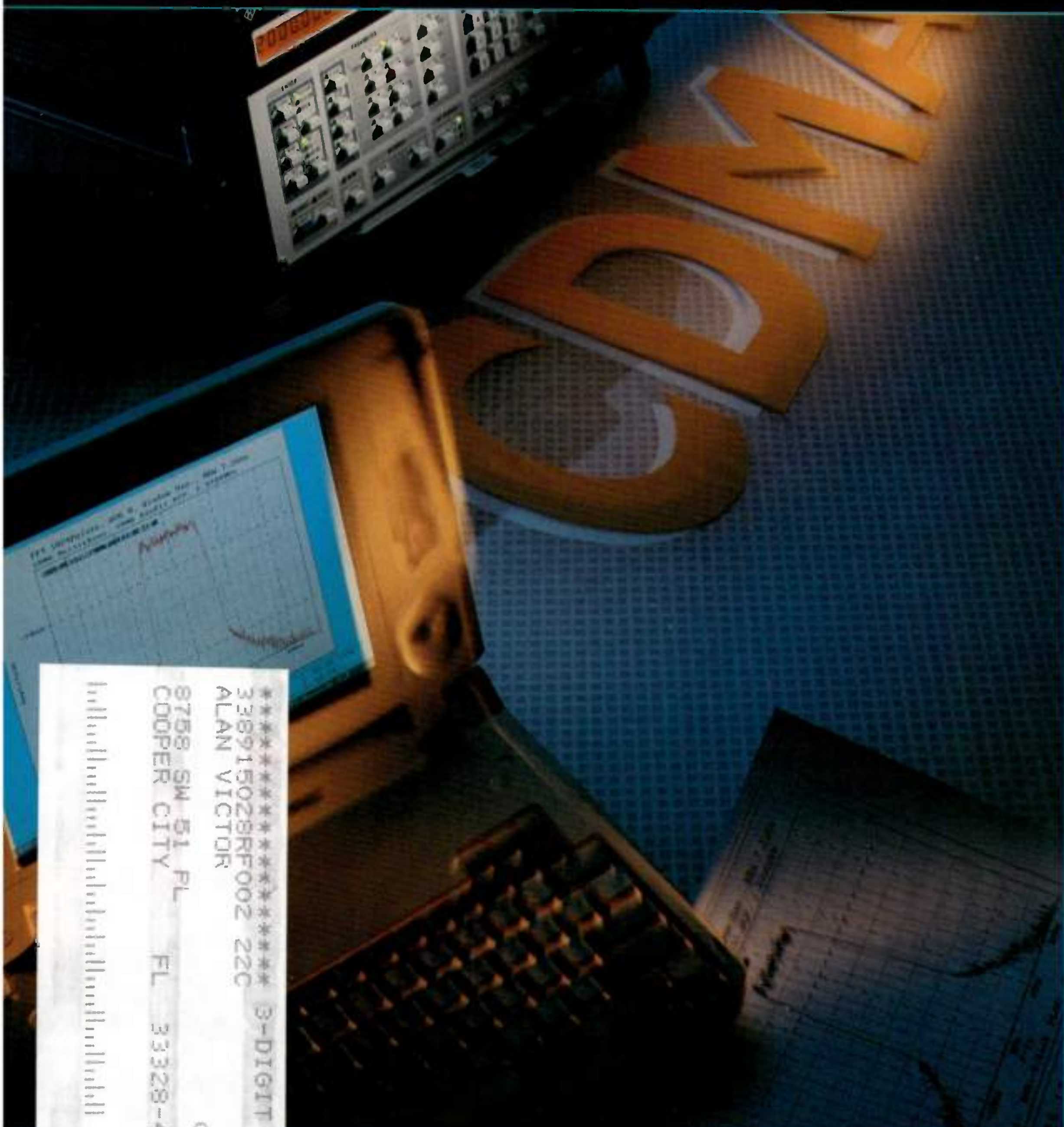
WIRELESS 17 YRS LEADERSHIP

RF design™

engineering principles and practices

July 1996

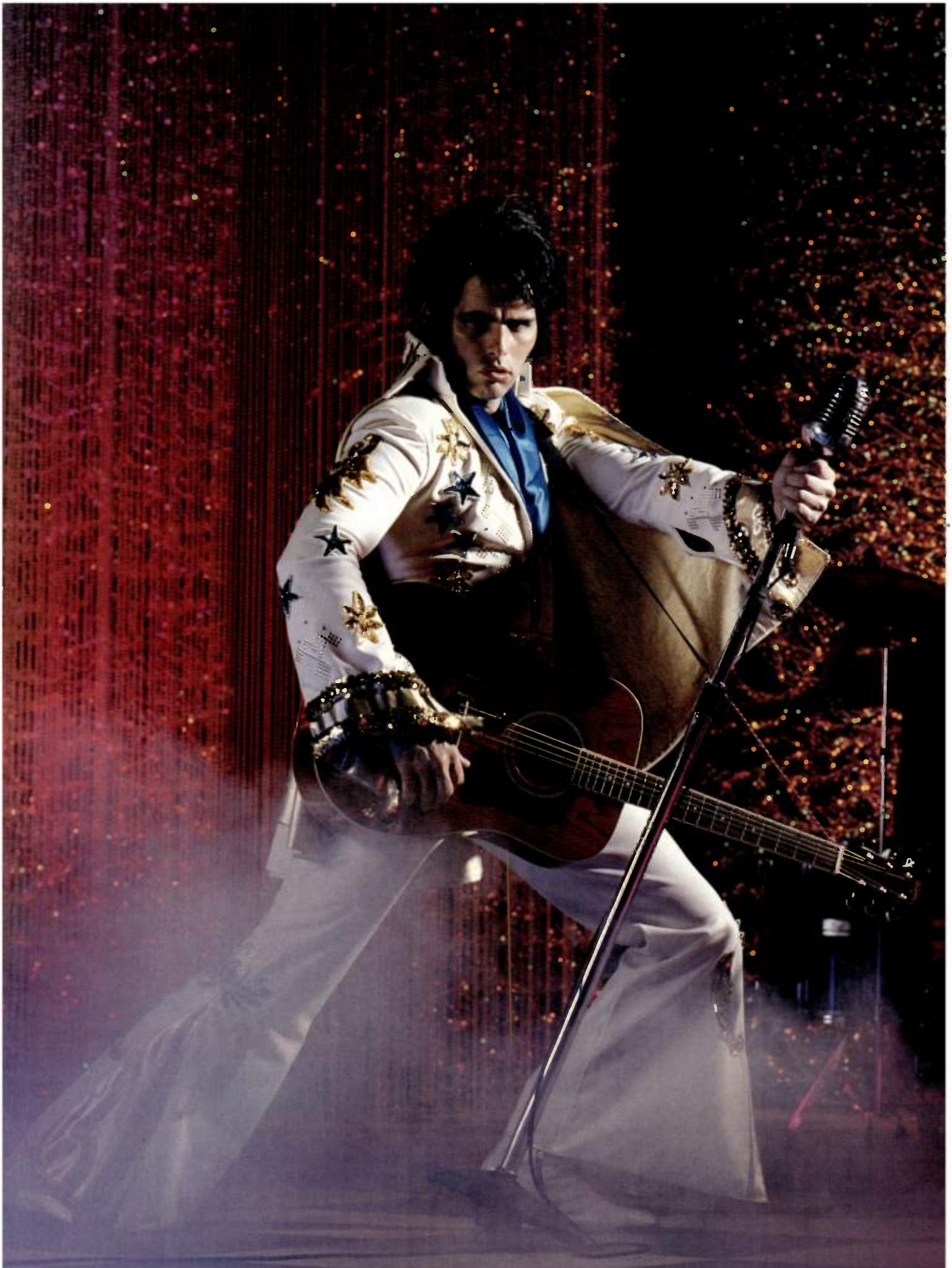


Featured Technology-

Power amplifiers

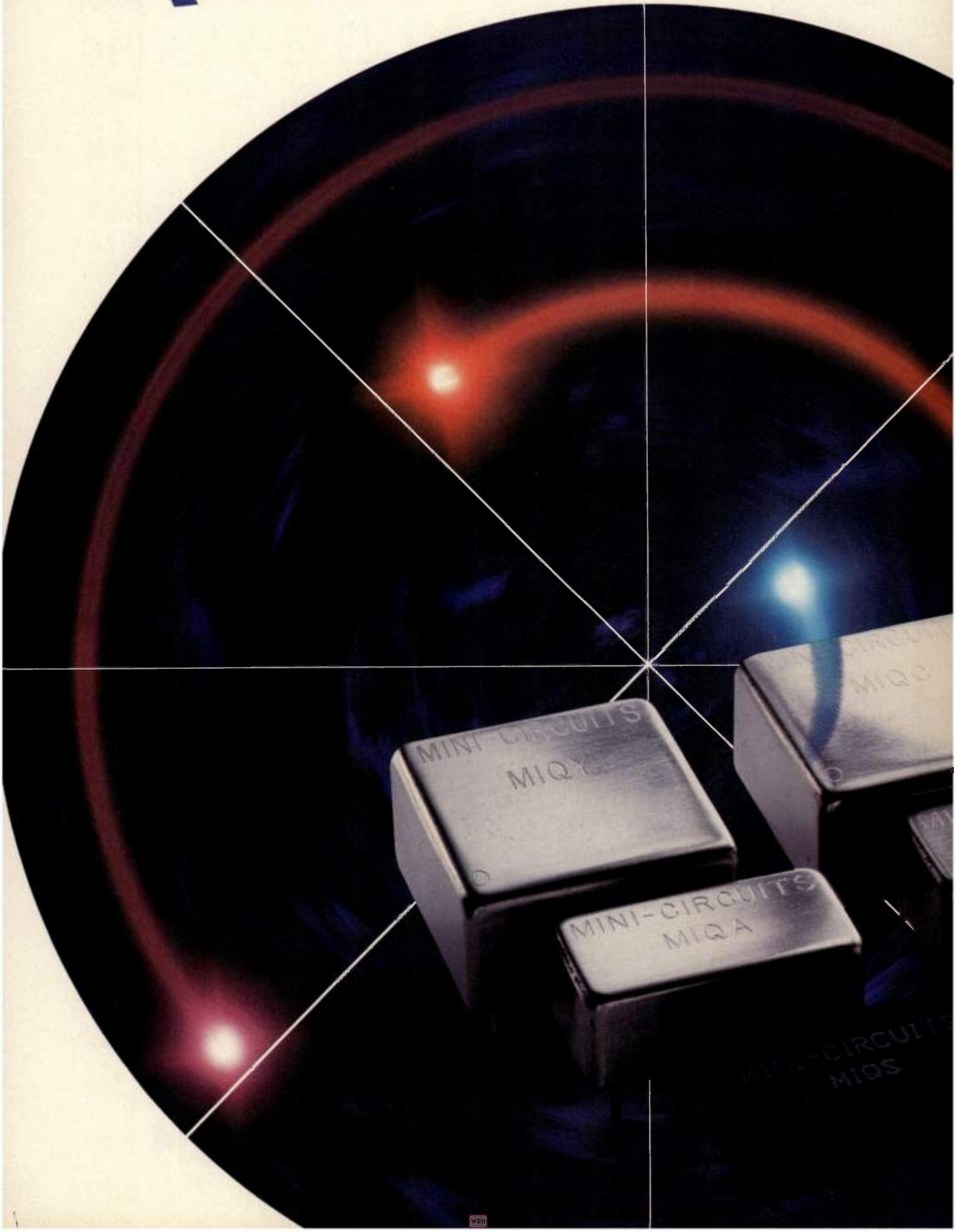
Tutorial-

Hints on cutting costs



Pictured above is Dale Sydnor, President of Noise Com, finally getting his "fifteen minutes of fame."

I&Q MODULATORS



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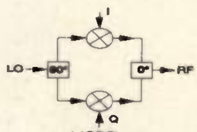

I/Q MODULATORS

MODEL	FREQ. (MHz) f_L	f_U	CONV. LOSS (dB) $\bar{x}$	σ	CARRIER REJ. (-dBc) Typ.	SIDEBAND REJ. (-dBc) Typ.	HARM. SUPPRESS (-dBc) Typ. 3xI/Q	5xI/Q	PRICE $ Qty. (1-9)
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MIQA-21M	20	23	6.2	0.14	50	40	48	65	39.95
MIQA-70M	66	73	6.2	0.10	38	38	48	58	39.95
MIQA-70ML	66	73	5.7	0.10	38	38	48	58	49.95
MIQA-91M	86	95	5.5	0.10	38	38	48	58	49.95
MIQA-100M	95	105	5.5	0.10	38	38	48	58	49.95
MIQA-108M	103	113	5.5	0.10	38	38	48	58	49.95
MIQA-195M	185	205	5.6	0.10	38	38	48	58	49.95
MIQC-38M	34	38	5.6	0.10	48	37	54	65	49.95
MIQC-88M	52	88	5.7	0.10	41	34	52	66	49.95
MIQC-176M	104	176	5.5	0.10	38	36	47	70	54.95
MIQC-895M	868	895	8.0	0.10	40	40	52	58	99.95
MIQC-1785M	1710	1785	9.0	0.30	35	35	40	65	99.95
MIQC-1880M	1805	1880	9.0	0.30	35	35	40	65	99.95
☐ MIQY-70M	67	73	5.8	0.20	40	36	47	60	19.95
☐ MIQY-140M	137	143	5.8	0.20	34	36	45	60	19.95
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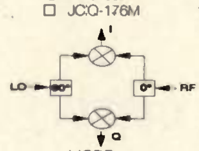
☐ JCIQ-88M	52	88	5.6	0.1	40	35	45	65	49.95
☐ JCIQ-176M	104	176	5.6	0.1	35	35	45	65	54.95



I/Q DEMODULATORS

MODEL	FREQ. (MHz) f_L	f_U	CONV. LOSS (dB) $\bar{x}$	σ	AMP. UNBAL. (dB) Typ.	PHASE UNBAL. (Deg.) Typ.	HARM. SUPPRESS (-dBc) Typ. 3xI/Q	5xI/Q	PRICE $ Qty. (1-9)
MIQA-10D	9	11	6.0	0.10	0.15	1.0	50	65	49.95
MIQA-21D	20	23	6.1	0.15	0.15	0.7	64	67	49.95
MIQA-70D	66	73	6.2	0.10	0.15	0.7	56	58	49.95
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☐ MIQY-70D	67	73	5.5	0.25	0.10	0.5	52	66	19.95
☐ MIQY-140D	137	143	5.5	0.25	0.10	0.5	47	70	19.95
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☐ JCIQ-176D	104	176	5.5	0.1	0.15	2	52	65	54.95
☐ JCIQ-895D	868	895	8.6	0.1	0.2	1	45	65	99.95
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Coming in August

- Antennas
- Coaxial cable tutorial
- Product focus: CAD for circuits

RF editorial

By Don Bishop
Editorial Director

MTT-S

No one can see all that there is to see at the Microwave Theory and Techniques Society's (MTT-S) International Microwave Symposium and Exhibition and the related symposia, conferences and seminars. Our technical editor, Gregg Miller, and I met many readers and visited a number of manufacturers. If we missed you, we'll look forward to seeing you at the next conference or at the *RF Design Seminar Series*, October 21–23 in Wakefield, MA.

My return from MTT-S coincides with my deadline for this column. News and product information derived from the show will follow in subsequent issues.

Editorial calendar

It's time to choose topics for the 1997 editorial calendar. The calendar sets the course for the magazine's editorial contents year-by-year, and suggestions for topics are gathered from readers, editorial advisory board members, marketing managers and the editors. *Your* suggestions are welcome. You can E-mail them to me at don_bishop@ intop.ccmail.compuserve.com, to Gregg at gregg_miller@intden.ccmail.compuserve.com or mail, phone or fax them, using our address and numbers in the masthead on page 10.

New team members

With this issue, we're welcoming our new associate editor, Patricia Werner, who comes to us with a background in technical writing, editing and desktop publishing. Along with coordinating the editorial side of the magazine production, Pat writes the news reports, tracks events and educational courses, and covers the publication of software and literature.

Another new member of the team is Ernest Worthman, our contributing editor. I came to know Ernest and to respect his work several years ago when he was the technical editor for *Communications* magazine, a competitor to one of the other magazines in the group to which *RF Design* belongs. With a B.S.E.E.T. and a background in hands-on technical work and in marketing, Ernest is a researcher and a technology enthusiast, who owns a company that trains people how to use computers. His first column appears in this issue.

Engineering and culture

The business of engineering electronic products crosses international frontiers and, at the same time, cultural boundaries. Someone I spoke with at MTT-S described one company's policy of exchanging engineers between its U.S. and Japanese facilities.

First, it was explained that more and more U.S.-made products are flowing to Japan, and the American engineers can bring their firsthand knowledge to help to resolve service problems. They learn more about the Japanese market, too.

Second, the Japanese engineers who work for a time in the U.S. facility are exposed to what some believe is a greater American propensity for individual creativity. The idea is that, when these engineers return home, they might help to infuse greater individual initiative to go along with already highly-developed teamwork.

I presume this policy works best within a multinational company—or are there other opportunities? If you or your company is involved in an exchange program of this type, I'd like to hear from you.

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"For example, our engineering and design capability enables us to respond rapidly to customer requirements. And with improved order handling and production capability, we are continually reducing delivery times."

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Colin I. W. Baxter, President
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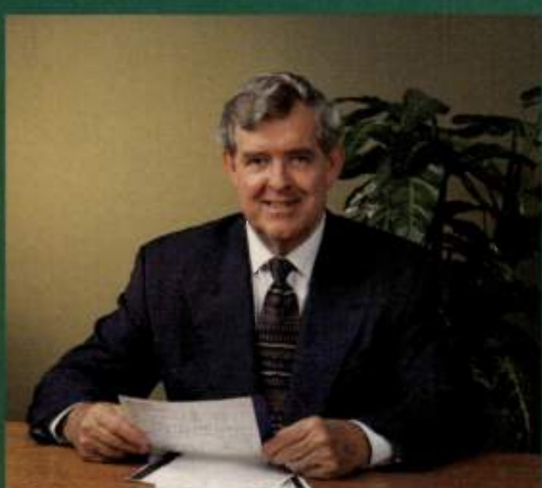

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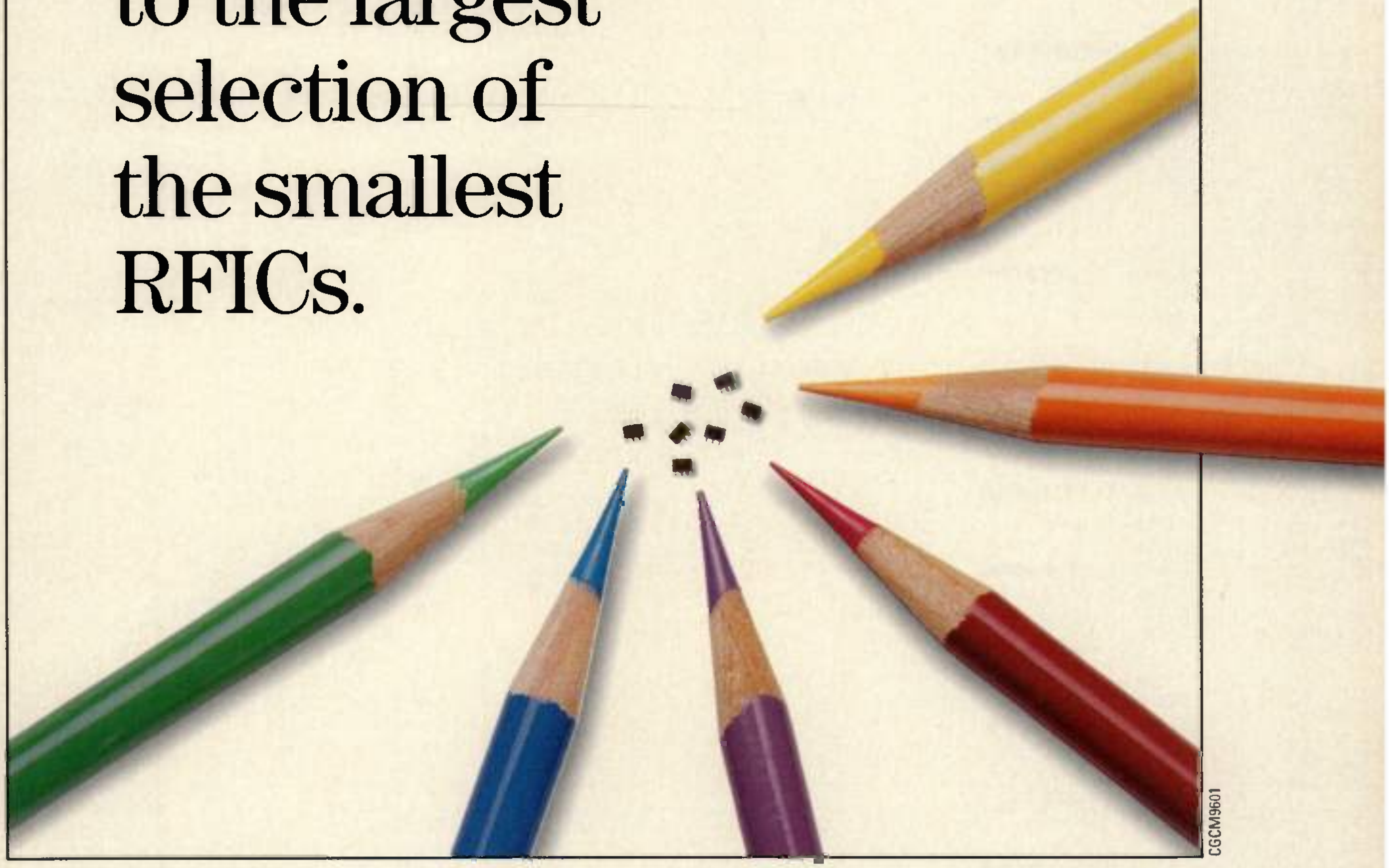

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INA-52063	Gain block	DC-1600	5	30	4.0	22	+15
▶ **MGA-81563**	Driver amp	100-6000	3	42	2.7	12	+27
▶ **MGA-82563**	Driver amp	100-6000	3	84	2.2	13	+31
MGA-86563	LNA	1500-6000	5	14	1.6	22	+15
MGA-87563	LNA	500-4000	3	4.5	1.6	14	+8

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2	TE5000	3	3.75	20	18.0	-	-	2	1.0	50	1800//+4
4	TE5010	3	3.75	30	14.0	-	-	3	2.0	60	1500//+3
6	TE5020	6	3.75	60	12.5	-	-	4	2.0	70	1500//+3
8	TE5030	6	3.75	60	10.0	90	12.5	5	2.0	80	1500//+3
2	TE5040	3	6.50	20	30.0	-	-	1	1.0	50	2700//0
4	TE5050	3	6.50	30	15.0	-	-	2	2.0	75	3100//0
6	TE5060	6	6.50	60	19.5	-	-	3	2.0	90	3100//0
8	TE5070	6	6.50	60	13.0	80	17.5	4	2.0	100	3100//0
2	TE5080	3	7.50	20	35.0	-	-	1	1.0	50	3000//0
4	TE5090	3	7.50	30	17.5	-	-	2	2.0	75	3300//0
6	TE5100	6	7.50	60	22.5	-	-	3	2.0	90	3300//0
8	TE5110	6	7.50	60	15.0	80	20.0	3	2.0	100	3300//0
2	TE5120	3	15.0	20	70.0	-	-	1	1.0	35	5000//-1
4	TE5130	3	15.0	30	35.0	-	-	2	2.0	60	5000//-1
6	TE5140	6	15.0	60	45.0	-	-	2	2.0	90	5000//-1
8	TE5150	6	15.0	60	30.0	80	40.0	3	2.0	100	5000//-1

21.4 MHz

NO. POLES	TEMEX P/N	PASSBAND dB	PASSBAND ±KHz	STOPBAND dB	STOPBAND ±KHz	STOPBAND dB	STOPBAND ±KHz	LOSS dB	RIPPLE dB-MAX	ULT. REJ. dB-MIN.	TERM. Ω/PF
2	TE5180	3	3.75	15	12.5	-	-	2	1.0	50	850//+6
4	TE5190	3	3.75	30	12.5	-	-	3	2.0	70	850//+5
6	TE5200	6	3.75	60	12.5	-	-	4	2.0	90	850//+5
8	TE5210	6	3.75	60	10.0	80	12.5	5	2.0	100	850//+5
2	TE5220	3	6.50	15	20.0	-	-	2	1.0	50	1300//+2
4	TE5230	3	6.50	30	22.5	-	-	3	2.0	70	1400//0
6	TE5240	6	6.50	60	22.5	-	-	4	2.0	90	1400//0
8	TE5250	6	6.50	60	17.5	80	22.5	4	2.0	100	1400//0
2	TE5260	3	7.50	15	25.0	-	-	2	1.0	50	1500//0
4	TE5270	3	7.50	30	25.0	-	-	3	2.0	70	1600//0
6	TE5280	6	7.50	60	25.0	-	-	4	2.0	90	1600//0
8	TE5290	6	7.50	60	20.0	80	25.0	4	2.0	100	1600//0
2	TE5300	3	15.0	15	50.0	-	-	2	1.0	45	3000//0
4	TE5310	3	15.0	30	45.0	-	-	3	2.0	60	3000//-1
6	TE5320	6	15.0	60	45.0	-	-	3	2.0	90	3000//-1
8	TE5330	6	15.0	60	33.0	80	45.0	4	2.0	100	3000//-1

45.0 MHz

NO. POLES	TEMEX P/N	MODE	PASSBAND dB	PASSBAND ±KHz	STOPBAND dB	STOPBAND ±KHz	LOSS dB	RIPPLE dB-MAX	ULT. REJ. dB-MIN.	TERM. Ω/PF
2	TE9420	3-OT	3	3.75	18	16.0	3	1	40	2000//-1.0
4	TE9310	3-OT	3	3.75	30	12.5	3	1	70	2000//-1.0
2	TE7420	3-OT	3	7.50	18	28.0	2	1	40	3000//-1.0
4	TE7430	3-OT	3	7.50	40	30.0	3	1	70	3000//-1.0
2	TE7440	3-OT	3	15.0	15	47.0	2	1	40	8000//-1.5
4	TE7450	3-OT	3	15.0	30	50.0	3	1	70	8000//-1.5
2	TE7730	FUND	3	15.0	15	50.0	2	1	40	1100//+1.5
4	TE7740	FUND	3	15.0	40	60.0	3	1	70	800//+1.0

70.0 MHz

NO. POLES	TEMEX P/N	MODE	PASSBAND dB	PASSBAND ±KHz	STOPBAND dB	STOPBAND ±KHz	STOPBAND dB	STOPBAND KHz	LOSS dB	RIPPLE dB-MAX	TERM. Ω/PF
2	TE10400	3-OT	3	7.5	18	30	35	-910	2	1	2000//-1
4	TE10410	3-OT	3	7.5	35	25	80	-910	3	1	2000//-1
2	TE10420	3-OT	3	10	15	30	35	-910	2	1	2500//-1
4	TE10430	3-OT	3	10	35	40	80	-910	3	1	2500//-1

90.0 MHz

NO. POLES	TEMEX P/N	MODE	PASSBAND dB	PASSBAND ±KHz	STOPBAND dB	STOPBAND ±KHz	STOPBAND dB	STOPBAND KHz	LOSS dB	RIPPLE dB-MAX	TERM. Ω/PF
2	TE10440	3-OT	3	7.5	18	30	35	-910	2	1	2000//-1
4	TE10450	3-OT	3	7.5	35	25	80	-910	3	1	2000//-1
2	TE10460	3-OT	3	10	15	30	35	-910	2	1	2500//-1
4	TE10470	3-OT	3	10	35	40	80	-910	3	1	2500//-1
4	TE10480	3-OT	3	15	30	50	80	-910	3	1	4000//-1

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Dear Editor,

The article by Glenn A. Parker that appears on pages 70–73 of your March 1996 issue concerns three-pole active filters. The author states that ". . . the symbolic solution for third-order filters is not documented." May I refer your readers to an article entitled "Design of Highpass, Lowpass and Bandpass Filters Using R-C Networks and Direct-current Amplifiers with Feedback" by C.C. Shumard of RCA Labs, published in *RCA Review*, pages 534–564, December 1950. In it, Shumard shows how to easily realize buffered, three-pole Butterworth response filters using only resistors and capacitors as the passive elements in configurations closely paralleling those shown by Parker.

Shumard was part of the TV development team headed by Vladimir K. Zworykin at RCA, where he first specialized in the design and development of wideband IF amplifiers for TV receivers. Please see "A Practical Television Receiver for the Amateur," *QST*, December 1938, pages 21–25, 72, 74 and 76.

Craig T. Brown, Inland Empire Labs
Corona, CA

Errata

We wish to correct an error in Bernard Kasmir's article, "Measuring Range and Reliability for Part 15 Systems," which appeared in our May 1996 issue. Table 1 on page 30 erroneous showed 99% reliability for 0 dB margi The table should have read 99% reli bility for 20 dB margin.

A table was omitted from page 84 Dr. Frederick H. Raab's article, "A Introduction to Class-F Power Amp fiers" in our May 1996 issue. Table appears below.

m	η: n = 1	n = 3	n = 5	n = ∞
1	1/2 = 0.500	9/16 = 0.563	75/128 = 0.586	$2/\pi$ = 0.637
2	2/3 = 0.667	3/4 = 0.750	25/32 = 0.781	$8/3\pi$ = 0.849
4	32/45 = 0.711	4/5 = 0.800	5/6 = 0.833	$128/45\pi$ = 0.905
∞	$\pi/4$ = 0.785	$9\pi/32$ = 0.884	$75\pi/256$ = 0.920	1 = 1.000

P_{max}: n = 1	n = 3	n = 5	n = ∞
1/8 = 0.125	9/64 = 0.141	75/512 = 0.146	$1/2\pi$ = 0.159

m = highest even harmonic properly controlled
n = highest odd harmonic properly controlled

Table 1. Characteristics with various combinations of harmonics.

INFO/CARD 35

RF calendar

July 21–26 **1996 IEEE AP-S International Symposium and URSI Radio Science Meeting**—*Baltimore*. Information: Jon Moellers, Steering Committee Chair, 445 Hoes Lane, P.O. Box 1331, Piscataway, NJ 08855-1331. Tel. 410-993-6774; Fax 410-993-7432.

August 14–16 **Piezoelectric Devices Conference**—*Kansas City, MO*. Information: Components Group, Electronic Industries Association, 2500 Wilson Blvd., Arlington, VA 22201-3834. Tel. 703-907-7500; Fax 703-907-7501.

19–21 **Wireless Communications Conference**—*Boulder, CO*. Information: Dr. Roger Marks, National Institute of Standards and Technology, 325 Broadway, MC 813.06, Boulder, CO 80303. Tel. 303-497-3037; Fax 303-497-7828; E-mail r.b.marks @ieee.org.

19–23 **IEEE International Symposium on Electromagnetic Compatibility**—*Santa Clara, CA*. Information: Gherry Pettit, Intel. Tel. 503-696-2994; Fax 503-640-6411.

September 8–12 **Electrical Overstress and Electrostatic Discharge Symposium**—*Orlando, FL*. Information: ESD Association, 7902 Turin Road, Suite 4, Rome, NY 13440-2069. Tel. 315-339-6937; Fax 315-339-6793; Web site http://www.eosesd.org.

16–20 **Accelerated Reliability Technology Symposium**—*Denver*. Information: Hobbs Engineering, 10218 Osceola Court, Westminster, CO 80030. Tel. 303-465-5988; Fax 303-469-4353; E-mail learn@hobbsengr.com.

17–20 **Electromagnetic Compatibility (EMC '96 Roma)**—*Rome*. Information: Prof. Mauro Feliziani, Dept. of Electrical Engineering, Univ. of Rome "La Sapienza," Via Eudossiana 18, 00184 Rome, Italy. Tel. +39 6 44585.809/ 44585.810; Fax +39 6 4883235/4825380; E-mail emc96rom@elettrica.ing.uniroma1.it.

19–20 **Electromagnetic Compatibility: Planning for Compliance in the U.S., Europe and Japan**—*Phoenix*. Information: Seminars Department, Underwriters Laboratories, 333 Pfingsten Road, Northbrook, IL 60062-2096. Tel. 847-272-8800 ext. 43481; Fax 847-509-6235; E-mail seminar@ul.com.

24–26 **Electrical Manufacturing & Coil Winding EMCW '96**—*Chicago*. Information: Electrical Manufacturing & Coil Winding '96, Dept. 77-5053, Chicago, IL 60678-5053. Tel. 708-260-9700 or 800-323-5155; Fax 708-260-0395.

25–26 **Chesapeake Electronics Show**—*Chantilly, VA*. Information: Bonnie Lasky, MACC, P.O. Box 513, Colmar, PA 18915. Tel. 215-822-6319; Fax 215-822-3332.

29–Oct. 2 **Wireless Workshop**—*Sedona, AZ*. Information: 100 S. Roosevelt Ave., Chandler, AZ 85226. Tel. 602-961-1382; Fax 602-961-4533.

October 7–10 **Signal Processing Applications and Technology**—*Boston*. Information: Megan Forrester c/o Miller Freeman, 600 Harrison St., San Francisco, CA 94107. Tel. 415-356-3391; Fax 415-905-2220; E-mail dsp@mfi.com.

7–11 **Wireless Technology '96**—*Providence*. Information: Dawn Averyt. Tel. 407-878-8200; Fax 407-879-7388; E-mail Expo96@aol.com.

8–10 **Microwaves & RF Conference and Exhibition**—*London*. Information: Beverley Lucan, Nexus Information Technology, Nexus House, Swanley, Kent, BR8 8HY, United Kingdom. Tel. +44-(0)1322-660070; Fax +44-(0)1322-661257.

21–23 ***RF Design* Seminar Series**—*Wakefield, MA*. Information: Intertec Presentations, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel. 303-220-0600; Fax 303-770-0253.

22–24 **Wescon '96 Technical Conference**—*Anaheim, CA*. Information: Wescon, 8110 Airport Blvd., Los Angeles, CA 90045. Tel. 800-877-2668 or 310-215-3976 ext. 251; Fax 310-641-5117; E-mail m.potthoff@ieee.org; Web site http://www.wescon.com.

29–30 **Radio Solutions '96**—*Birmingham, UK*. Information: LPRA Secretariat, Walker Mitchell Ltd., Brearley Hall, Luddenden Foot, Halifax HX2 6HS, United Kingdom. Tel. and fax +44 (0) 1422 88 69 50.

29–31 **Signal Processing Applications and Technology**—*Santa Clara, CA*. Information: DSP Associates, 49 River St., Waltham, MA 02154. Tel. 617-891-6000; Fax 617-899-4449; E-mail icspat@dspnet.com.

RF courses

Training Program for Cellular, PCS Staff—Home-study course. Information: Virginia Polytechnic Institute and State University, Mobile & Portable Radio Research Group, 840 University City Blvd., Pointe West Commons, Suite 1, Blacksburg, VA 24061-0350. Tel. 540-231-2970; Fax 540-231-2968.

Wireless Networks and Mobile Communications—July 30–Aug. 2, 1996, Washington, DC. Information: Learning Tree International, Reston Town Center, 1805 Library St., Reston, VA 22090-9919. Tel. 800-843-8733; Fax 800-709-6405; E-mail uscourses@learningtree.com.

Communications Tech Training—1996 schedule for Orland Park, IL, Aug. 6–8, Sept. 3–5, Oct. 8–10, Nov. 5–7, Dec 3–5, Information: Andrew Corp., Dept. 355, P.O. Box 9000, San Fernando, CA 91341-9978. Tel. 800-255-1479 ext. 117.

George Washington University—Washington, DC. *Cellular and Wireless Telephony*—Aug. 12–16; *Wireless Infrastructure Network Engineering for Cellular, PCS, LE, and WPBX*—Oct. 21–25. Information: George Washington University, Continuing Engineering Education, Academic Center, Room T-308, 801 22nd St. N.W., Washington, DC. 20052. Tel. 202-994-6106 or 800-424-9773; Fax 202-872-0645; E-mail ceepinfo@ceep.vpaa.gwu.edu.

1996 CEI-Europe in Cambridge, United Kingdom
Wireless Digital Communications: Mobile, Cellular, Personal, Voice and Data Networks—Sept. 30–Oct. 4; *Applied RF Techniques: Linear Circuits*—Sept. 30–Oct. 4; *Adaptive Synchronous Receiver Structures for Mobile Communications*—Sept. 30–Oct. 4.
in Baveno, Italy—
Mobile and Wireless Personal Communications Networks —Oct. 14–18; *Modern Digital Modulation Techniques*—Oct. 14–18; *Modeling and Simulation of Communication Systems*—Oct. 15–18; *Speech and Channel Coding for Mobile Communication*—Oct. 21–23; *Digital Cellular and PCS Communications: The Radio Interface*—Oct. 21–25; *Spread-Spectrum and CDMA*—Oct. 21–25.
in Barcelona, Spain—
Mobile Cellular and PCS Telecommunications Systems—Nov. 11–13; *RF Circuit Components*—Nov. 18–21; *Receivers and Transmitters*—Nov. 18–22; *Advanced Digital Receivers for Wireless Communications*—Nov. 18–21; *Cellular and Personal Communications Infrastructure*—Nov. 20–22. Information: CEI-Europe, P.O. Box 910, S-612 25 Finspong, Sweden. Tel. 46 122 175 70; Fax 46 122 143 47; E-mail cei.europe@one.se.

DSP Without Tears—Aug. 26–28, Washington, DC; Sept. 30–Oct. 2, San Jose, CA. Information: Z Domain Technologies, 325 Pine Isle Court, Alpharetta, GA 30302. Tel. 800-967-5034 or 770-587-4812; Fax 770-518-8368; E-mail dsp@mindspring.com.

Celwave University—Marlboro, NJ. Modules offered include: *Antenna Basics, Advanced Theory, Towertop Amplifiers, Bidirectional Amplifiers, Filters* and *Combiners*. Information: Gail Magid, Celwave, 2 Ryan Road, Marlboro, NJ 07746-1899. Tel. Sales Engineering Dept. 800-235-9283.

Georgia Tech Continuing Education
Near-field Antenna Measurements and Microwave Holography—Sept. 10–13, Boulder, CO; *Wireless Communication*—Sept. 24–27, Atlanta, GA. Information: Dept. of Continuing Education, Georgia Institute of Technology, Atlanta, GA 30332-0385. Tel. 404-894-2547; E-mail conted@gatech.edu;Web http://www.conted.gatech.edu.

Johns Hopkins University Whiting School of Engineering
—courses in Washington, DC
Modern Terrestrial Radio Systems Engineering—Sept. 9–12; *Cellular and Personal Communications Services (PCS): Systems Engineering and Perspective with Emphasis on CDMA Technology*—Sept. 16–19; *Wireless and Personal Communications Systems*—Oct. 7–9; *Wireless Digital Communications Systems: Specification, Test and Evaluation*—Oct. 21–23; *The Telecommunications Revolution and its Impact on Organizational Planning*—Oct. 28–30. Information: OEI 14515 Barkwood Drive, Rockville, MD 20853. Fax 301-871-4942; E-mail info.oei@apl.jhu.edu.

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RF Input-Main PLL RF Input-Aux PLL	2.5GHz 510MHz	2.0GHz 510MHz	1.2GHz 510MHz	1.1GHz 1.1GHz	2.0GHz 1.1GHz	550MHz 550MHz
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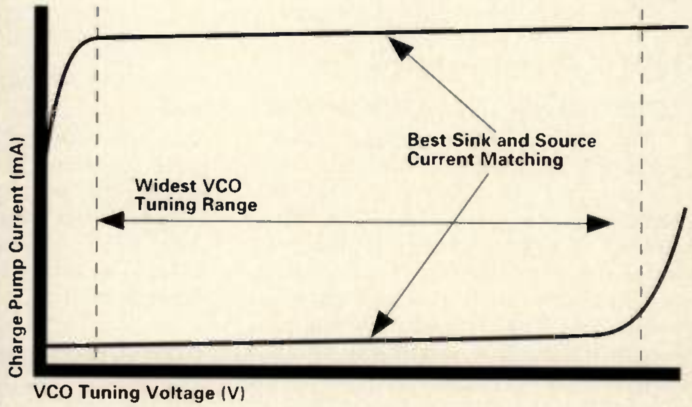

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INFO/CARD 92

NVLAP completes international standards

The National Voluntary Laboratory Accreditation Program (NVLAP) at the Commerce Department's National Institute of Standards and Technology (NIST) is now fully compatible with standards for laboratory accreditation and quality system management used worldwide. The completion of the project will bring the traceability of U.S.-made measurements in line with international practices. This will facilitate broad acceptance of valid calibration and test results and avoid barriers between international trading partners.

Among other fields of testing, NVLAP accredits testing laboratories in the areas of electromagnetic compatibility, telecommunications and ionizing radiation dosimetry. Calibration fields for laboratory accreditation include electromagnetic-RF/microwave, ionizing radiation, and time and frequency.

The new NVLAP 1996 directory lists approximately 700 domestic and foreign testing and calibration laboratories accredited by NVLAP to meet the international standards requirements as of January 1996. For a copy of the NVLAP 1996 Directory (SP-810), send a self-addressed mailing label to Laboratory Accreditation Program, Bldg. 820, Room 282, NIST, Gaithersburg, MD 20899-0001. Tel. 301-975-4016; Fax 301-926-2884; E-mail nvlap@enh.nist.gov.

Nonstop toll collection uses electronic system

Amtech Systems will install the first nonstop electronic toll collection (ETC) system to be implemented in Brazil for the renovation of the Rio-Niterói bridge. The ETC system will keep traffic moving smoothly as toll collection goes into effect on the bridge.

A small electronic Amtech tag placed on each vehicle's windshield will register toll transactions electronically via sensors in toll lanes. The toll amount is then billed to the patron's account or credit card.

The Rio-Niterói Bridge is a large hub in the road system of southeastern Brazil and connects the cities of Rio de Janeiro and Niterói, with more than 80,000 vehicles traversing the five-mile causeway each day.

Amtech Systems provides wireless identification, tracking and monitoring technologies for the intelligent transportation industry.

Contract:

SpaceCom, Redsat and TeleComm form Strategic Alliance—SpaceCom has joined with Redsat of Mexico and TeleComm de Mexico to market SpaceCom's proprietary FM Cubed satellite transmission technology and equipment in Mexico, Central America, the Caribbean and South America.

The new service, called InfoSat Plus will provide point-to-multipoint satellite distribution of audio and data services via the Solidaridad satellite. RedSat will oversee the marketing activities while SpaceCom will market the service.

Business Briefs

Retlif Expands Test Site—Retlif Testing Laboratories, a strategic compliance organization, has acquired a new one-third acre site with a newly retrofitted 4,000 sq. ft. building including 12,500 sq. ft. of laboratory space. The addtional land will allow Retlif to offer a full 10-meter test site. The move was a result of Retlif's growing volume of test services, particularly in the area of environmental simulation.

Wavetek and Yokogawa Electric Form Alliance—Yokogawa Electric Corporation of Japan, which manufactures industrial automation and test and measurement equipment, will distribute Wavetek's products in Japan under the name Yokogawa-Wavetek. The two companies will collaborate on new products for the communications test market worldwide.

Microwave Online Services Established—To bring the RF-microwave industry on line, Microwave Online Services in Boulder, CO, develops and maintains Web sites focused exclusively on the industry. The company combines engineering, marketing, software development, graphic design and management of the RF Globalnet, the engineering and business center of the RF-microwave world.

Stanford Telecom Microwave Systems Division Wins Contract from Rockwell International—The contract is for the first phase of design and production of a DDS-based synthesizer for the single-channel, anti-jam, man-portable terminal (SCAMP) program. The total contract is valued at $5.9 Million.

Global Positioning Satellite (GPS) market—KW Microwave has created a new division to serve the growing GPS market. The military retrofitting of aircraft fleet, the opening of GPS satellites to commercial aircraft and the acceptance of GPS products for almost anything that moves have created a demand for GPS products.

Hewlett-Packard Research and Development Moves In-house—The company is focusing resources on improving the HP High-frequency Structure Simulator (HFSS). HFSS is an electromagnetic tool that computes S-parameters for passive, 3D structures. Hewlett-Packard cites long simulation times and large computer memory requirements as the main roadblocks to widespread use of HFSS among RF and microwave circuit designers.

Tessco Technologies Acquires Cartwright Communications—Tessco Technologies, Sparks, MD, supplies more than 14,000 products from over 230 major manufacturers to the cellular telephone, paging, PCS and mobile radio-dispatch markets. Cincinnati-based Cartwright Communications, a value-added distributor of 4,000 radio communications equipment products, will operate as a wholly-owned subsidiary of Tessco.

DESIGN FROM START-TO-ART

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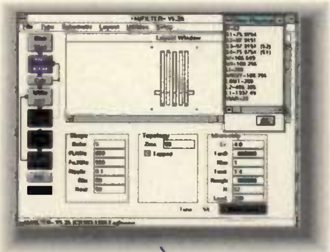
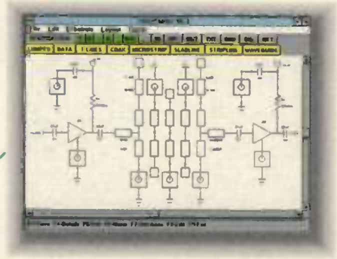
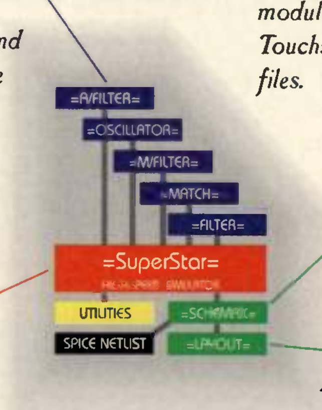
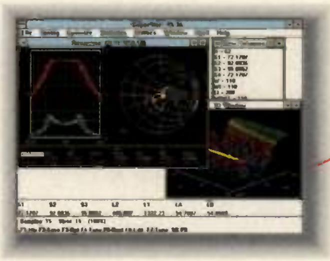


2 Schematics make it easy to enter your own designs, modify synthesis created designs and merge subcircuits. Schematics and simulation are integrated for back annotation. The GENESYS schematic module also writes Touchstone and Spice files.







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INFO/CARD 16

Wideband high-efficiency power amplifiers

By Robert J. Dehoney

An important characteristic of power amplifiers is the efficiency of their output stages. This is particularly true for solid-state amplifiers designed for military or heavy-duty commercial use. This use imposes the conflicting requirements of low device junction temperature to insure reliability and operation at high ambient temperatures.

Inefficient operation affects the entire power amplifier package, requiring not only complex cooling schemes for the output devices, but requiring a larger power supply and, for portable applications, a larger generator as well, just to supply the extra power that is wasted as heat.

An attractive approach to minimizing the loss in the output devices is to operate them in as close to a *switch mode* as possible. [1,2]

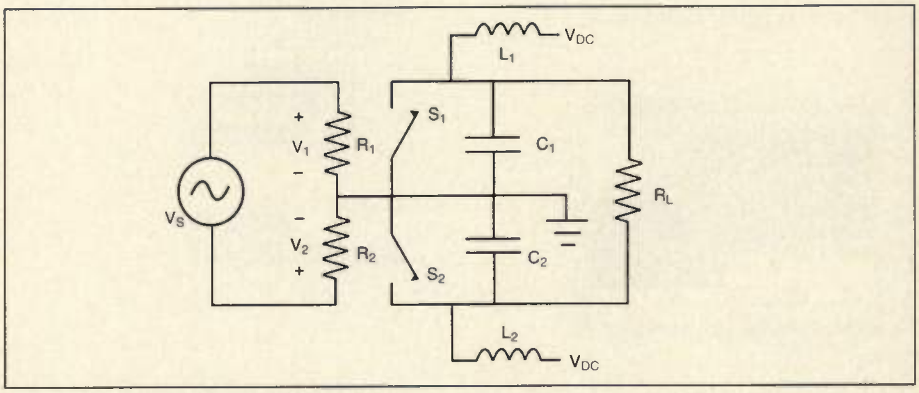

Figure 1. Push-pull output stage circuit.

Unfortunately, switch mode operation requires that the harmonic voltages and currents be closely controlled. If reactive (reflective) filters are used, reactive elements must be located (electrically) near the output devices. In some cases, this cannot be done. For example, in a 2-to-30 MHz, 5 kW amplifier, 16 power modules might be required, combined in a four-level binary combiner. Considering cabling and combiner electrical length, there can be many electrical degrees at the harmonic frequencies between the harmonic filter assembly and the device junctions. At some frequencies, this can cause severe loss of efficiency and large voltage swings at the device junction, which can damage or destroy the devices.

To explore this situation, we will simulate some idealized circuits to see what happens when a switched source operates in a narrowband load.

Figure 1 shows a PSPICE version of a push-pull output stage with output devices simulated by capacity-loaded switches. In a real-world circuit, of course, there would be many other circuit elements, such as DC blocking capacitors, biasing circuits and a baanced-to-unbalanced transformer (balun). In fact, this could be one of many modules driven

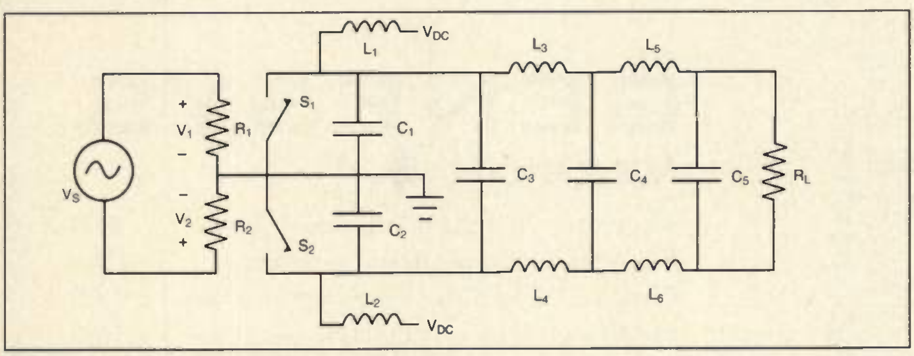

Figure 2. Power amplifier with a harmonic filter.

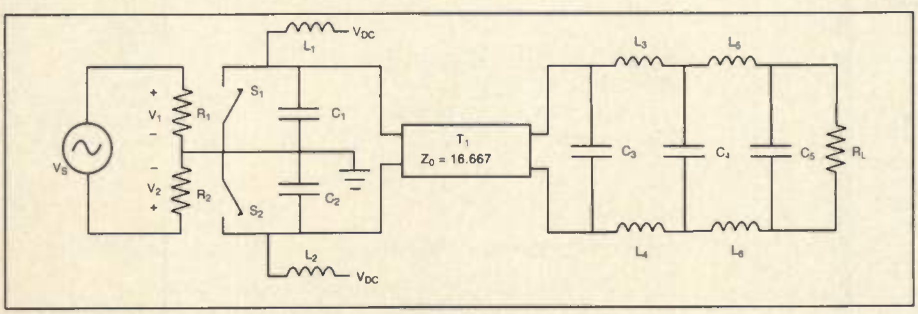

Figure 3. Power amplifier of Figure 2 with a transmission line.

T1 Wavelengths at Fundamental	Harmonic Number	Harmonic Amplitude (V)	Fundamental Amplitude (V)	Idc (A)	Eff
0.04	7th	162	65.2	6	76%
0.058	5th	365	57.7	5.6	63%
0.108	3rd	574	41.3	3.9	46%

Table 1. PSPICE statistics for Figure 3.

in parallel and working into a combiner. However, the effects we wish to explore are not influenced by these elements; they only clutter up the simulation.

The switches are operated by the voltages across R_1 and R_2, V_1 and V_2. When V_1 exceeds 0.01 V, S_1 closes and looks like a low resistance. When V_1 is less than 0 V, S_1 opens and looks like a high resistance; V_2 similarly controls S_2. We arbitrarily choose V_s to be a 1 MHz, 10 V peak sine wave and C_1 and C_2 to be 100 pF. We chose the *on* resistance, R_S, to be 1 Ω and the *off* resistance to be 1 MΩ. For reasons that will be evident later, we choose $R_L = 16.667\ \Omega$.

With L_1 and L_2 large enough to give a substantially constant current, the output across R_L is a square wave of about four times V_{DC} peak-to-peak. With V_{DC} = 28 V, PSPICE gives the spectral components and average current as follows:

I_{DC} = 5.42 A

HARMONIC NO.	AMPLITUDE (V)
Fundamental	57.50
2	0.53
3	19.20
4	0.54
5	11.50
6	0.54
7	8.20
8	0.55
9	6.30

The efficiency is the ratio of the fundamental power divided by the DC power. For this idealized circuit, the inductor current is:

$$I_L = \frac{2V_{DC}}{R_L + 4R_S} \tag{1}$$

The peak-to-peak square wave output is:

$$V_{PP} = \frac{4V_{DC}R_L}{R_L + 4R_S} \tag{2}$$

The efficiency is:

$$Eff = \frac{(0.8116)R_L}{R_L + 4R_S} \tag{3}$$

As R_S approaches 0, the efficiency approaches 81.16%. For our circuit, the efficiency is 65.3%.

We now investigate what happens

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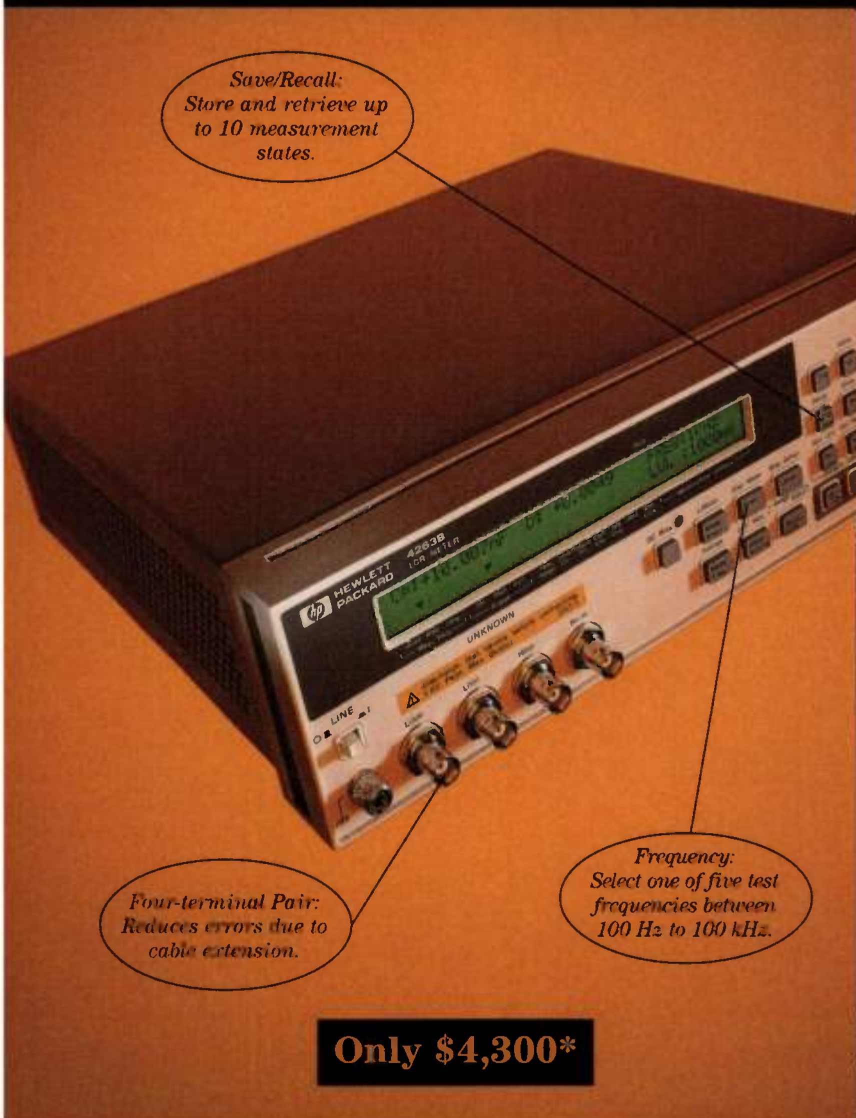

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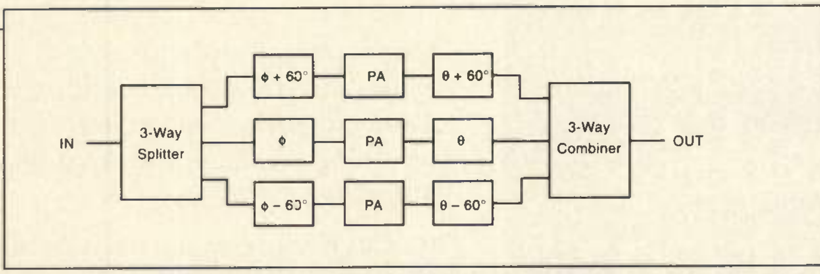

Figure 4. Block diagram of an amplifierdesigned to eliminate the effects of harmonic power.

when the harmonic power is not absorbed, but is instead reflected back into the switches. This is the case when a harmonic filter is added to our simple power amplifier, as shown in Figure 2.

C_3–C_5 and L_3–L_6 form a balanced version of a five-pole Chebychev lowpass filter. This one is chosen to have 0.01 dB ripple. PSPICE gives the spectral results and average current as follows:

I_{DC} = 6.208 A

HARMONIC NUMBER	ACROSS C3 (V)	ACROSS RL (V)
Fundamental	69.91	67.01
2	0.47	0.24
3	9.58	0.78
4	0.16	0.16
5	3.29	0.15
6	0.42	0.17
7	1.74	0.10
8	0.21	0.35
9	1.41	0.20

Compared to the no-filter case, the results are gratifying. The efficiency is up to 77.5%, and the harmonic power has not caused any great voltage swings on the switches. However, let us now move further into the real world. Our 2 to 30 MHz amplifier will need eight switched harmonic lowpass filters. We cannot filter each of the 16 modules before combining. Controlling the phase would be horrendous. We have to do our combining before we get to the switched filter assembly.

We simulate this situation by adding a 16.667 Ω transmission line between the switches and the filter, as shown in Figure 3).

Table 1 summarizes a lot of PSPICE running time and shows in a dramatic fashion the problems harmonic power can cause.

The line lengths transform the reactance of the filter into an inductance that resonates with the switch capacity at the harmonic frequency. The stronger the harmonic, the more voltage is generated, and the more the efficiency is degraded. Note also the dramatic reduction in fundamental power. Even if the devices could tolerate the huge voltages, the transmitter is forced to operate at as much as 4 dB below its capability.

How can this effect be avoided?

One solution is to use non-reflecting filters. These are usually designed as highpass and lowpass pairs called *complimentary filters*. The highpass section is terminated and serves only to absorb the harmonic energy. However, some problems are associated with the design of these filters [3]. Another consideration is the number of additional circuit components required. To maintain a low voltage standing wave ratio (VSWR), our eight filters might need nine reactive elements each. The complimentary sections might need seven elements each, adding 56 elements to our design.

The *other* approach is the subject of the following section. Consider the block diagram of Figure 4. The phase-shifting sections provide the indicated phase shift over the band F_{low} to $5{\times}F_{high}$. Table 2 traces the phase shift through the diagram.

The first phase-shift section shifts the reference channel (B) by a variable amount X and shifts the A and

CHANNEL	HARMONIC	RELATIVE PHASE INTO PAs (°)	RELATIVE PHASE OUT OF PAs (°)	PHASE INTO COMBINER (°)
	Fundamental	60	60	0
	3rd	n/a	180	120
A	5th	n/a	300	240
	7th	n/a	60	0
	9th	n/a	180	120
	Fundamental	0	0	0
	3rd	n/a	0	0
B	5th	n/a	0	0
	7th	n/a	0	0
	9th	n/a	0	0
	Fundamental	−60	−60	0
	3rd	n/a	−180	−120
C	5th	n/a	−300	−240
	7th	n/a	−60	0
	9th	n/a	−180	−120

Table 2. Results of phase shift used in Figure 4.

HARMONIC NO.	FREQUENCY (Hz)	FOURIER COMPONENT	NORMALIZED COMPONENT	PHASE (°)	NORMALIZED PHASE (°)
1	3.00×10^{6}	1.269×10^{1}	1.00	-7.259×10^{1}	0.00
2	6.00×10^{6}	1.091×10^{-1}	8.592×10^{-3}	9.776×10^{1}	1.703×10^{2}
3	9.00×10^{6}	3.779	2.977×10^{-1}	1.562×10^{2}	2.288×10^{2}
4	1.20×10^{7}	2.145×10^{-1}	1.690×10^{-2}	-6.668	6.592×10^{1}
5	1.50×10^{7}	3.482	2.743×10^{-1}	1.474×10^{1}	8.732×10^{1}
6	1.80×10^{7}	2.314×10^{-1}	1.823×10^{-2}	-1.553×10^{2}	-8.273×10^{1}
7	2.10×10^{7}	1.474	1.161×10^{-1}	-1.308×10^{2}	-5.822×10^{1}
8	2.40×10^{7}	2.358×10^{-1}	1.857×10^{-2}	6.867×10^{1}	1.413×10^{2}
9	2.70×10^{7}	2.145	1.689×10^{-1}	5.756×10^{1}	1.301×10^{2}

Table 3(a). Fourier components of transient response V(7,8) with a DC component of -7.294504×10^{-2}.

HARMONIC NO.	FREQUENCY (Hz)	FOURIER COMPONENT	NORMALIZED COMPONENT	PHASE (°)	NORMALIZED PHASE (°)
1	3.00×10^{6}	1.273×10^{1}	1.000	-1.321×10^{2}	0.000
2	6.00×10^{6}	1.980×10^{-1}	1.555×10^{-2}	2.811×10^{1}	1.602×10^{2}
3	9.00×10^{6}	3.858	3.031×10^{-1}	-5.122×10^{1}	8.086×10^{1}
4	1.20×10^{7}	1.310×10^{-1}	1.029×10^{-2}	5.796×10^{1}	1.901×10^{2}
5	1.50×10^{7}	1.893	1.487×10^{-1}	5.472×10^{1}	1.868×10^{2}
6	1.80×10^{7}	1.850×10^{-1}	1.453×10^{-2}	-1.689×10^{2}	-3.683×10^{1}
7	2.10×10^{7}	1.776	1.395×10^{-1}	-1.482×10^{2}	2.803×10^{2}
8	2.40×10^{7}	2.131×10^{-1}	1.674×10^{-2}	-6.007×10^{1}	7.202×10^{1}
9	2.70×10^{7}	1.845	1.449×10^{-1}	-7.645×10^{1}	5.564×10^{1}

Table 3(b). Fourier components of transient response V(9,10) with a DC component of -9.80847×10^{-2}.

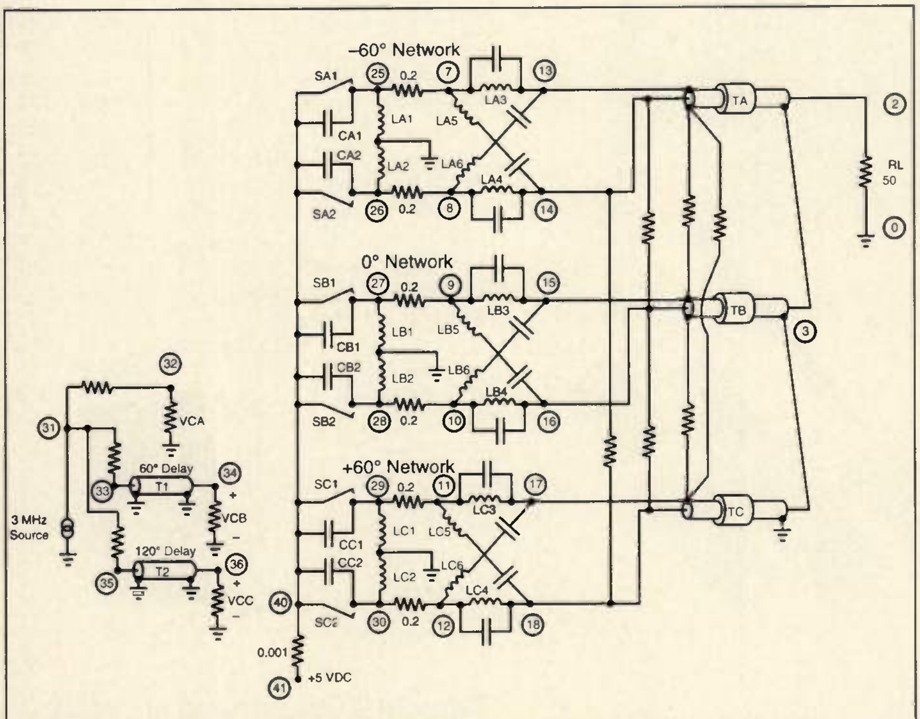

Figure 5. Circuit designed to test whether phase-shifting eliminated the effects of harmonic power.

C channels by X+60 degrees and X–60 degrees as shown. We assume the signal is sinusoidal at this point in the amplifier, so the phase shift of the harmonics is of no interest. We make the assumption that the power amplifiers (PAs) generate square waves in phase with their inputs. The phases shown are with respect to the reference channel phases. Note that in any channel, the shift of the Nth harmonic is N times the shift of the fundamental.

Each channel goes through a last phase shifter as shown. Here, the reference channel is shifted by a variable angle Y, and the other two channels are shifted by Y–60 degrees and Y+60 degrees.

The three channels are summed in the output combiner, with the fundamental and seventh harmonic power adding in the output load and the other harmonic power being dissipated in the combiner dump ports.

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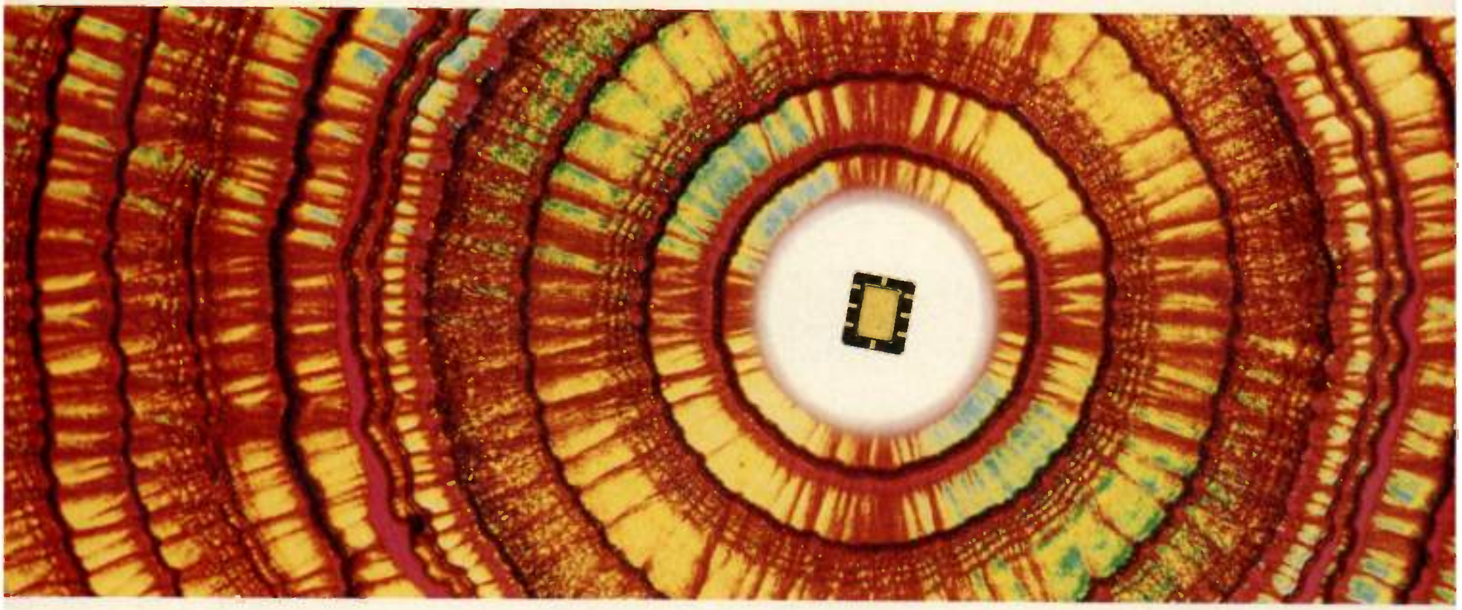

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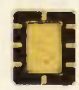

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PSPICE file for Figure 5

```
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   FROM DELAY LINE.
.TRAN .5US 4.6US 4US .5US UIC
.FOUR 3MEG V(7,8) V(9,10) V(11,12)
   V(2,0) V(40,41)
.OPTIONS ITL5=0
V1 31 0 SIN (0 200 3MEG)
R1 31 32 50
R4 32 0 50
R2 31 33 50
T1 33 0 34 0 Z0=50 F=3MEG
   NL=.166667
R5 34 0 50
R3 31 35 50
T2 35 0 36 0 Z0=50 F=3MEG
   NL=.3333333
R6 36 0 50
SA1 25 40 32 0 SMOD
SA2 26 40 0 32 SMOD
CA1 25 40 10PF
CA2 26 40 10PF
LA1 25 0 1MH IC=.6
LA2 26 0 1MH IC=.6
RA3 7 25 .2
RA4 8 26 .2
SB1 27 40 34 0 SMOD
SB2 28 40 0 34 SMOD
CB1 27 40 10PF
CB2 28 40 10PF
LB1 27 0 1MH IC=.6
LB2 28 0 1MH IC=.6
RB3 27 9 .2
RB4 28 10 .2
SC1 29 40 36 0 SMOD
SC2 30 40 0 36 SMOD
CC1 29 40 10PF
CC2 30 40 10PF
LC1 29 0 1MH IC=.6
LC2 30 0 1MH IC=.6
RC3 29 11 .2
RC4 30 12 .2
VDC 41 0 DC 5
RR 41 40 .001
.MODEL SMOD VSWITCH (RON=.1
   ROFF=1MEG VON=.01 VOFF=0)
RX1 13 15 25
RX2 13 17 25
RX3 15 17 25
RY1 14 16 25
RY2 14 18 25
RY3 16 18 25
LA3 7 13 3.380367UH
LA4 8 14 3.380367UH
CA3 7 13 808.4901PF
CA4 8 14 808.4901PF
LA5 7 19 .2254798UH
LA6 8 20 .2254798UH
CA5 20 13 12.12079NF
CA6 19 14 12.12079NF
LB3 9 15 1.253415UH
LB4 10 16 1.253415UH
CB3 9 15 320.7782PF
CB4 10 16 320.7782PF
LB5 9 21 89.46184NH
LB6 10 22 89.46184NH
CB5 22 15 4.494299NF
CB6 21 16 4.494299NF
LC3 11 17 .4973075UH
LC4 12 18 .4973075UH
CC3 11 17 118.9422PF
CC4 12 18 118.9422PF
LC5 11 23 33.17181NH
LC6 12 24 33.17181NH
CC5 23 18 1.783167NF
CC6 24 17 1.783167NF
TA 13 14 1 2 Z0=16.7 F=21MEG NL=.1
TB 15 16 3 1 Z0=16.7 F=21MEG NL=.1
TC 17 18 0 3 Z0=16.7 F=21MEG NL=.1
RL 2 0 50
.END
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.01–500	AU-1310	30	0.5	2.0:1	1.3	1.4	1.5	8	$325
.01–500	AU-1332	45	0.5	2.0:1	1.3	1.4	1.5	10	$350
.01–1000	AM-1300	25	0.75	2.0:1	1.4	1.6	1.8	6	$375
1–100	AU-2A-0110	30	0.5	2.0:1	1.2	1.2	1.3	3	$275
1–100	AU-3A-0110	50	0.5	2.0:1	1.2	1.2	1.3	10	$300
1–500	AU-1A-0150	14	0.5	2.0:1	2.7	2.8	2.9	10	$200
1–500	AU-2A-0150	30	0.5	2.0:1	1.3	1.4	1.5	8	$275
1–500	AU-3A-0150	45	0.5	2.0:1	1.3	1.4	1.5	10	$300
1–500	AU-4A-0150	60	0.5	2.0:1	1.3	1.4	1.5	10	$325
1–1000	AM-2A-000110	25	0.75	2.0:1	1.4	1.6	1.8	8	$300
1–1000	AM-3A-000110	35	0.75	2.0:1	1.4	1.6	1.8	9	$350
20–200	AU-1158	30	0.5	2.0:1	2.7	2.7	2.7	17	$275
50–90	AU-1001	14	0.25	1.3:1	5.0	5.0	5.0	18	$200
50–90	AU-2A-1158	30	0.25	1.3:1	2.7	2.7	2.7	20	$275
50–90	AU-3A-1263	43	0.25	1.3:1	1.5	1.5	1.5	20	$325
50–350	AU-1210	18	0.5	2.0:1	2.6	2.7	2.8	10	$200
100–1000	AM-1331	35	0.75	2.0:1	1.4	1.6	1.8	15	$400
100–2000	AMMIC-1348	14	1	2.2:1	5.0	4.3	4.6	14	$395
100–2000	AMMIC-1318	6	1	2.2:1	4.5	4.0	4.0	12	$350
500–1000	AMMIC-1141	10	0.5	1.5:1	6.0	6.0	6.0	10	$200
500–1000	AM-2A-0510	24	0.5	2.0:1	1.4	1.5	1.6	0	$300
500–1000	AM-3A-0510	37	0.5	2.0:1	1.4	1.5	1.6	9	$350
500–1500	AM-3A-0515	30	0.5	2.0:1	1.5	1.8	2.2	4	$375
500–2000	AM-2A-0520	19	0.75	2.0:1	1.4	1.9	2.4	-4	$350
500–2000	AM-3A-0520	30	0.75	2.0:1	1.4	1.9	2.4	5	$400
500–2000	AM-4A-0520	40	1	2.0:1	1.4	1.9	2.4	5	$450
1000–2000	AM-2A-1020	19	0.5	2.0:1	1.8	2.1	2.4	4	$325
1000–2000	AM-3A-1020	30	0.5	2.0:1	1.8	2.1	2.4	10	$375
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HARMONIC NO.	FREQUENCY (Hz)	FOURIER COMPONENT	NORMALIZED COMPONENT	PHASE (°)	NORMALIZED PHASE (°)
1	3.00×10^{6}	1.257×10^{1}	1.000	1.686×10^{2}	0.000
2	6.00×10^{6}	1.202×10^{-1}	9.564×10^{-3}	5.940×10^{1}	-1.092×10^{2}
3	9.00×10^{6}	4.993	3.972×10^{-1}	1.452×10^{2}	-2.336×10^{1}
4	1.20×10^{7}	1.109×10^{-1}	8.824×10^{-3}	1.527×10^{1}	-1.533×10^{2}
5	1.50×10^{7}	2.983	2.373×10^{-1}	1.045×10^{2}	-6.409×10^{1}
6	1.80×10^{7}	9.917×10^{-2}	7.890×10^{-3}	1.038×10^{2}	-6.477×10^{1}
7	2.10×10^{7}	2.122	1.688×10^{-1}	8.811×10^{1}	-8.050×10^{1}
8	2.40×10^{7}	1.897×10^{-1}	1.509×10^{-2}	2.495×10^{1}	-1.437×10^{2}
9	2.70×10^{7}	4.346×10^{-1}	3.458×10^{-2}	1.774×10^{1}	-1.509×10^{2}

Table 3(c). Fourier components of transient response V(11,12) with a DC component of 7.503634×10^{-2}.

This scheme substantially does what we want. The only potential problem is the seventh harmonic, but it is not strong to begin with, and in a real transmitter, insertion loss and filter Qs minimize its build-up.

To test this scheme, a PSPICE file was generated for the circuit shown in Figure 5. In this circuit, three push-pull switched sources drive three networks, which shift the phase as shown. Their outputs are combined in a tri-hybrid. The switches are operated by a 3 MHz source with outputs at 0°, –60°, and –120°. The simulation used delay lines to get the phase shift, because only one frequency was being run. In a real amplifier, constant phase-shifting sections would be used. To minimize running time, only a single two-pole circuit was used in each channel. This limited the bandwidth ratio but still allowed fair accuracy up to the ninth harmonic.

The test was intended to answer the following questions:

a) What is the harmonic level at the switches? Is the square wave well-preserved?

b) What is the harmonic level at the output? Is it sufficiently low so that reflected harmonic power is of no consequence?

c) What is the conversion efficiency, DC to fundamental power?

Table 3 is taken from the PSPICE printout. V(7,8), V(9,10), and V(11,12) are the spectral components at the switches and show fairly well behaved harmonics.

V(2,0) is the output showing a fundamental of 38 V peak with minuscule third and fifth harmonic and a seventh harmonic more than 16 dB down.

V(40,41) shows a 3.58 A DC current, giving an input DC power of 5(3.58) = 17.9 W. The fundamental power is 38(38)/2/50 = 14.44 W giving 80.6% efficiency.

HARMONIC NO.	FREQUENCY (Hz)	FOURIER COMPONENT	NORMALIZED COMPONENT	PHASE (°)	NORMALIZED PHASE (°)
1	3.00×10^{6}	3.804×10^{1}	1.000	-7.627×10^{1}	0.000
2	6.00×10^{6}	9.542×10^{-2}	2.508×10^{-3}	-6.220×10^{1}	1.407×10^{1}
3	9.00×10^{6}	1.084	2.851×10^{-2}	1.746×10^{2}	2.509×10^{2}
4	1.20×10^{7}	1.065×10^{-1}	2.800×10^{-3}	-1.430×10^{2}	-6.676×10^{1}
5	1.50×10^{7}	4.592×10^{-1}	1.207×10^{-2}	-1.077×10^{2}	-3.140×10^{1}
6	1.80×10^{7}	1.275×10^{-1}	3.350×10^{-3}	6.862×10^{1}	1.449×10^{2}
7	2.10×10^{7}	5.569	1.464×10^{-1}	6.060×10^{1}	1.369×10^{1}
8	2.40×10^{7}	6.798×10^{-2}	1.787×10^{-3}	-1.609×10^{2}	-8.468×10^{1}
9	2.70×10^{7}	1.201	3.158×10^{-2}	-1.545×10^{2}	-7.826×10^{1}

Table 3(d). Fourier components of transient response V(2,0) with a DC component of 1.832173×10^{-2}.

We can conclude that the scheme works. If we can design networks that provide relative phase of –60°, 0°, and 60° over a broadband, we can implement the blocks of Figure 4 and cancel, or at least considerably reduce, third and fifth harmonics.

How do we design the networks? There may be an analytic way to synthesize them, but I'm not aware of it. One way that works reasonably well is to find the pole frequencies for a two-channel, 120° network, then add in the reference channel using approximate geometric mean pole frequencies. An example will make this clear. Assume we want to build a 2-to-30 MHz amplifier. The output constant-phase allpass networks will then have to cover 2 to 150 MHz to pass the fifth harmonic.

Using an allpass network program such as *RF Design's* RFD–0193, we find that 4 poles/channel will give us 120° with less than 0.2° error over the band. The pole frequencies are:

CHANNEL A (MHz)	CHANNEL C (MHz)
100.7689000	625.233000
22.0397700	58.933720
5.0904600	13.611760
0.4798198	2.977109

We find the geometric mean of each pair as trial pole frequencies for channel B. This gives us:

CHANNEL B (MHz)
251.0060600
36.0400560
8.3240690
1.1951886

The phase shift produced in each channel is:

$$\arctan\left(\frac{P_1}{F}\right)+\arctan\left(\frac{P_2}{F}\right)+\arctan\left(\frac{P_3}{F}\right)+\arctan\left(\frac{P_4}{F}\right) \qquad (4)$$

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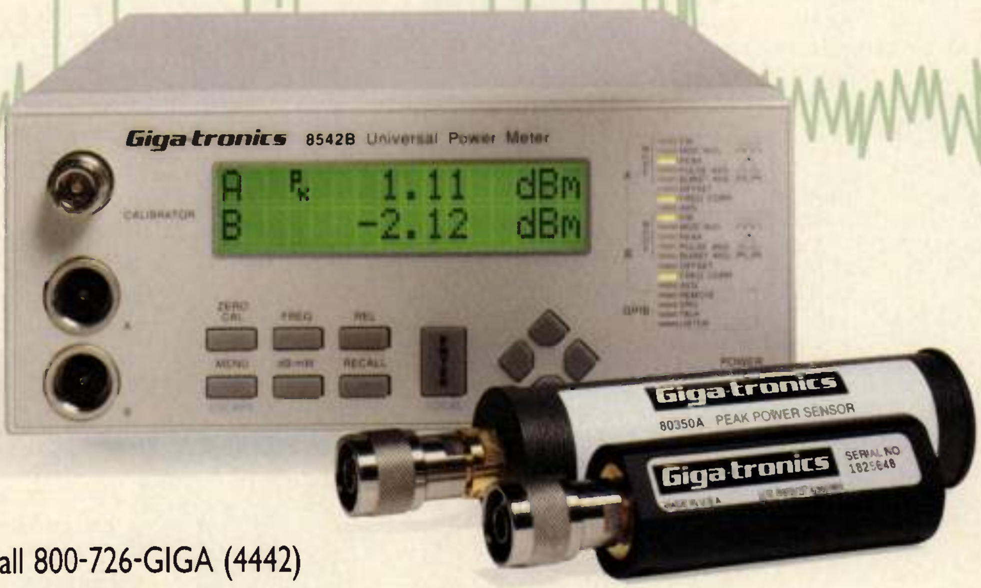

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1	3.00×10^{6}	1.261×10^{-7}	1.000	1.799×10^{2}	0.000
2	6.00×10^{6}	3.881×10^{-8}	3.078×10^{-1}	-1.178×10^{2}	-2.978×10^{2}
3	9.00×10^{6}	4.103×10^{-8}	3.254×10^{-1}	-1.778×10^{2}	-3.578×10^{2}
4	1.20×10^{7}	3.287×10^{-8}	2.607×10^{-1}	-1.694×10^{2}	-3.493×10^{2}
5	1.50×10^{7}	2.452×10^{-8}	1.945×10^{-1}	-1.675×10^{2}	-3.474×10^{2}
6	1.80×10^{7}	2.299×10^{-8}	1.823×10^{-1}	1.795×10^{2}	-4.701×10^{-1}
7	2.10×10^{7}	1.187×10^{-8}	1.496×10^{-1}	-1.662×10^{2}	-3.462×10^{2}
8	2.40×10^{7}	2.155×10^{-8}	1.709×10^{-1}	1.176×10^{2}	-6.238×10^{1}
9	2.70×10^{7}	1.540×10^{-8}	1.221×10^{-1}	-1.633×10^{2}	-3.433×10^{2}

Table 3(e). Fourier components of transient response V(41,40) with a DC component of 3.582645×10^{-3}.

where P_1–P_4 are the pole frequencies in that channel, and F is the frequency.

Subtracting the shift in channel A from the shift in channel B, we find that the difference ranges from 55°–65° over the 2-to-150 MHz band. In an effort to reduce the error, we slightly shift the pole frequencies of channel B. We find that reducing P_1 and P_2 by 0.933 and increasing P_3 and P_4 by the same factor gives an error of less than 1° over the band. This is certainly accurate enough for our purposes. With the pole frequencies established, we can easily obtain the circuit values for our choice of network.

Conclusion

A scheme has been presented that enables switch-mode operation of output devices over a broad bandwidth without tuning. A method has been shown for designing the critical phase shifting circuits RF

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About the Author

Robert Dehoney received his BSEE from MIT in 1950. After graduation, he worked for Allen B. DuMont Labs designing RF equipment for UHF TV. He retired from Fairchild Weston in 1987 as a Technical Director in ECM. He is now a private consultant. He can be reached at 4602 Palm Boulevard, Isle of Palms, SC 29451, or by phone at 803-886-5785.

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By Mark Atkinson
Orion Software

With the rapidly expanding market for broadband, spurious-free receivers and transmitters, the need for determining internally generated spurious signals has grown. There are several potential sources for spurious signals in modern communication equipment that need to be addressed. With the advent of microprocessors, digital signal processors (DSPs) and multiple-conversion high-frequency designs, the possibility for problems with spurious frequencies is great. This is especially true in broadband receivers with sub-microvolt sensitivity and transmitters with tight inband spurious output specifications such as cable television (CATV) products including modulators, block converters, and processors (frequency translation).

Good shielding and filtering of input and output (I/O) lines can virtually eliminate the microprocessor and DSP as a source, but the conversion process in communication devices will generate internal spurious frequencies that will show up in the input band or intermediate frequency (IF) of receivers and transmitters. Because of the non-linear characteristics of mixers and the harmonics from oscillators, the potential exists for the large number of internally generated spurious signals that appear in the input range of receivers and the output of transmitters and other signal sources.

As an example, consider a dual-conversion design (Figure 1), which has internally generated spurious inband signals. The local oscillator (Lo) outputs can be passed through appropriate lowpass or bandpass filters to virtually eliminate harmonics of the oscillators reaching the mixer. However, the mixer creates products of its own. In addition, stray coupling on the printed circuit board, inadequate shielding and coupling between mixers are the main sources for signals getting through to the mixers.

Depending on the design objective, many of the higher-order (M & N) spurious frequencies can be ignored, as usually the higher M and N values are the weaker spurious frequencies. If the design is for a sensitive receiver, or other high-specification product requiring a clean input or output, then an analysis should be performed. This analysis can be time-consuming, but using a computer program to perform the analysis speeds the process.

One particular program identifies spurs in single-, double-, and triple-conversion designs. It also selects the best IF and fixed Lo frequencies in dual conversion for minimum spurs by searching through selected ranges. The program runs under Windows and is easy to use. The output includes tabular values and spectrum plots to show where spurious frequencies occur. To identify potentially stronger spurs, any M or N of 2 or lower is highlighted in red both in the table and spectrum plot. A pull-down menu allows for complete variability in selecting the design to evaluate. The harmonics

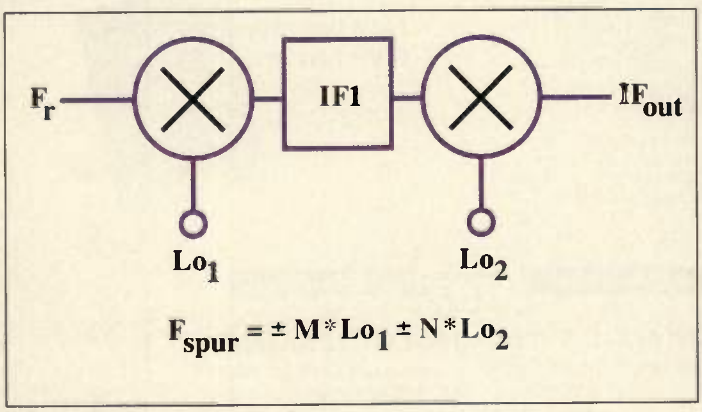

Figure 1. Dual-conversion receiver.

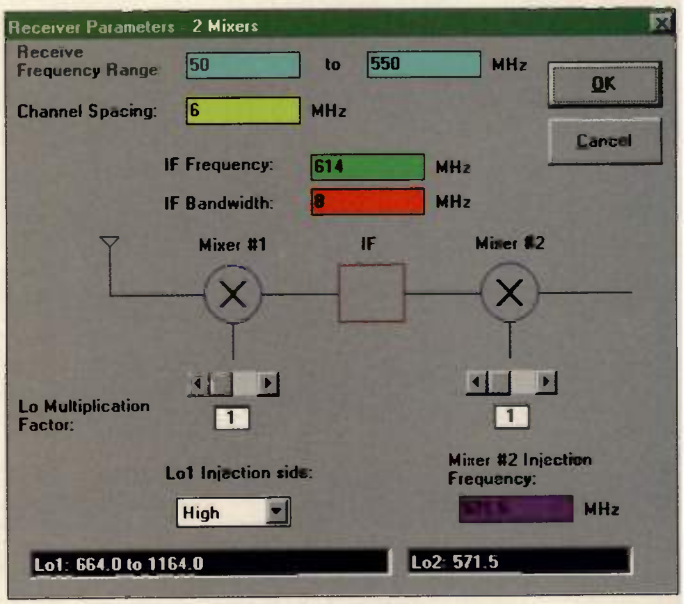

Figure 2. Dual-conversion parameter window.

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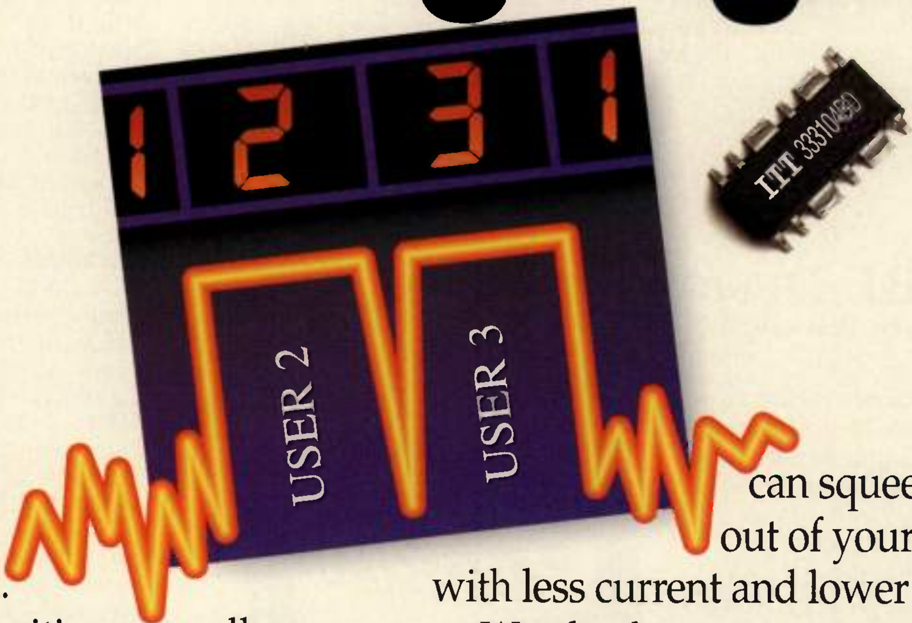

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of the oscillator can be selected from 1 to 99 in the options menu, with the default value of 20. The table is correlated with the spectral line plots, and selecting any line in the spur table will turn the associated spectral line green in the spectral display. Calculations can be performed in kHz, MHz, and GHz, as set by the user.

Parameters are entered through graphical parameter windows that vary depending on the type of design considered. Suppose one selects *double conversion*. A window appears with a block diagram of a double-conversion receiver. The values are entered for the parameters as shown in Figure 2, (*page 36*). With the use of the tab key or mouse, one can move through the parameter window. It is possible to select low- or high-side injection, and the multiplication factor of each Lo from 1 to 12. The values of the Los based on input frequency and IF appear automatically at the bottom of the parameters window. The functionality of the program can best be demonstrated with several examples.

Example 1: CATV demodulator

A CATV demodulator or channel processor input section typically has an input range from 50 to 550 MHz and a second IF output that matches the standard TV IF of 43.5 MHz. A dual-conversion approach is normally used. In this example, a first IF is selected to be 614 MHz with 8 MHz bandwidth. Initially, high-side injection is selected, which shows the Lo, a voltage-controlled oscillator (VCO), to cover from 664 to 1164 MHz. The second Lo is also selected to be high-side to convert down the frequency. After all the values are entered, including a channel step size of 6 MHz, click on "OK." A table window appears with all the spurs presented, clearly showing how they were determined.

The M and N values, spur values, and the two oscillators are presented in a table as shown in Figure 3. The spectral line plot at the bottom of the screen offers a graphical plot of the spurs. The amplitude is based on the M and N value with the lowest values of M and N having the highest amplitude in the plot. An M or N that equals 2 or less is highlighted in red along with the corresponding spectral line in the spectral plot.

The frequency axis of the spectrum plot has a range that includes the desired range of 50–550 MHz plus one half of the IF bandwidth added onto each end; therefore, spurs could occur from 46 to 554 MHz (as is shown) because of the 8 MHz bandwidth of the first IF filter. For simplicity, the IF bandwidth filter response is assumed to be a brick wall. The table shows spurs in the desired input range. It is now possible to select low-side in-

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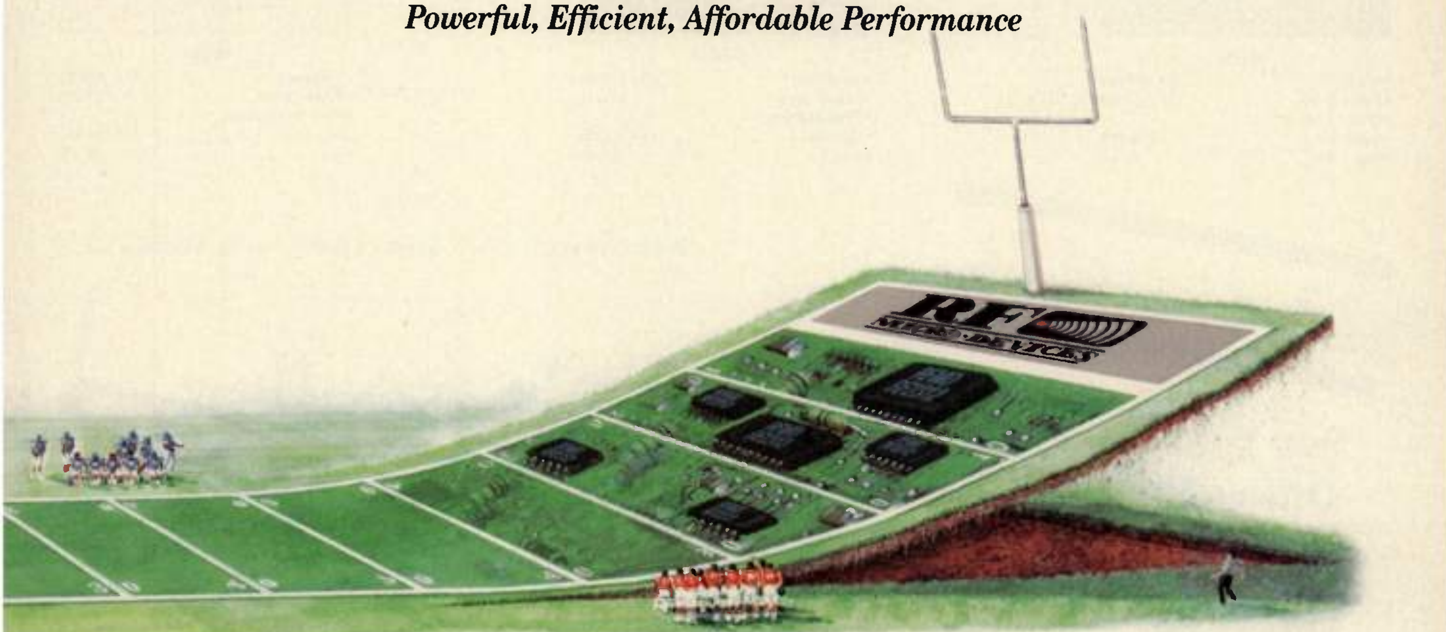

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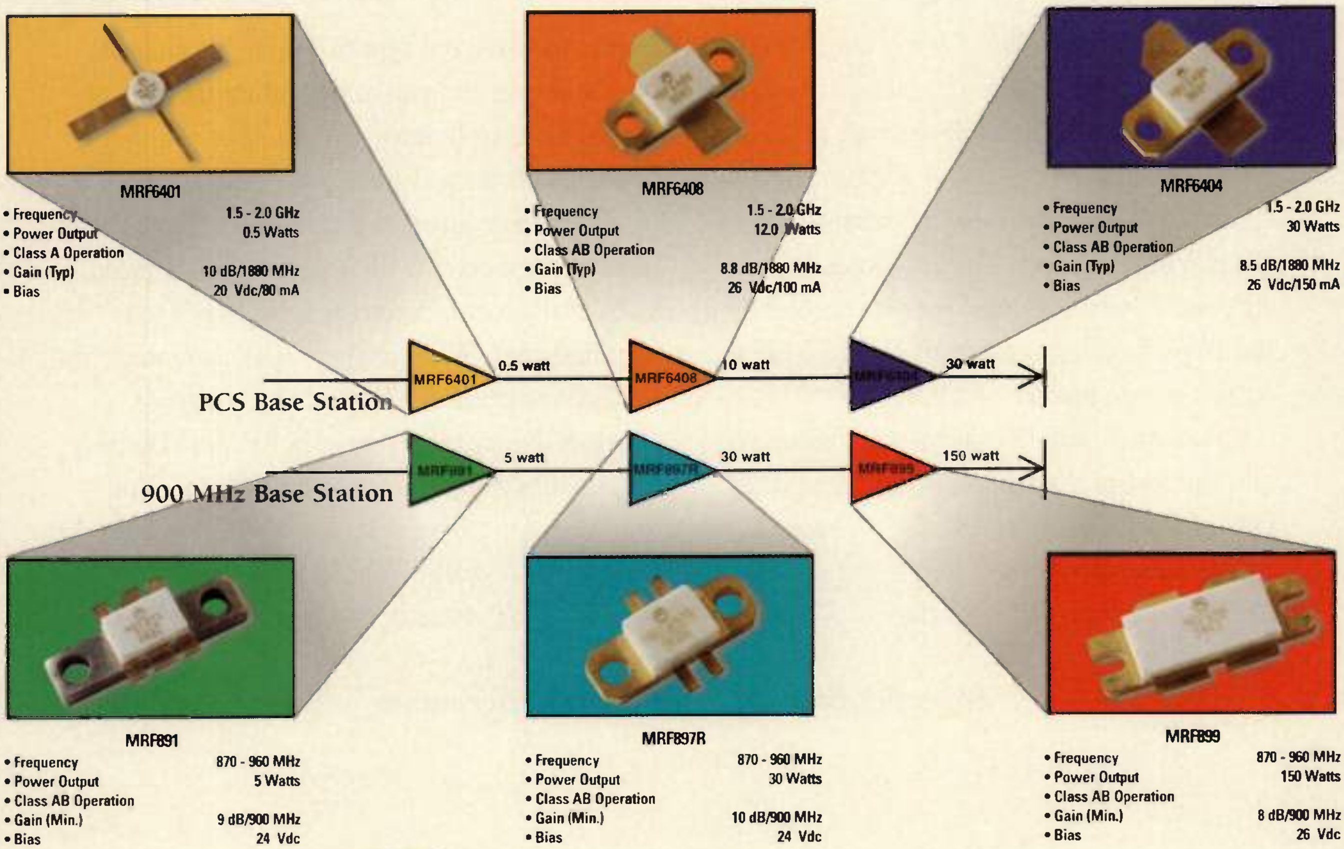

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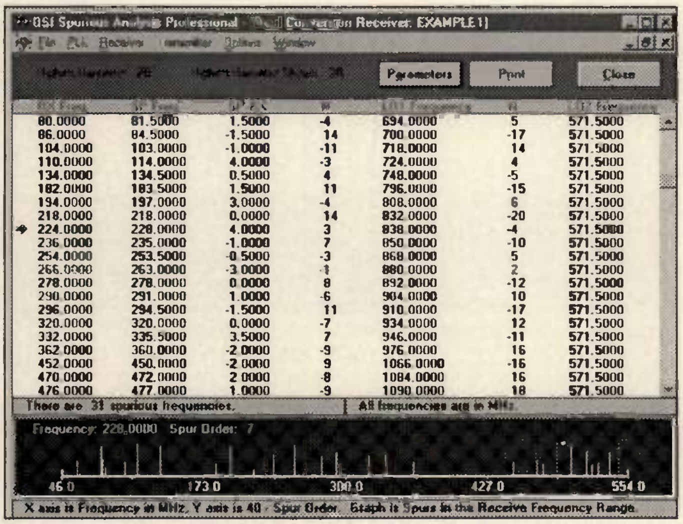
jection and different first IF values to reduce the number of spurs. A search feature calculates spur values based on an automatic search of a range of IF and fixed Lo values. Although the ideal result is free of spurs, one could shift the worst spurs by massaging the intermediate and local oscillator frequencies from a more trouble-some spur to a benign spot in the band. It is easy to modify parameters and recalculate. From the table window, click on "Parameters," and



80.0000	81.5000	1.5000	-4	694.0000	5	571.5000
86.0000	84.5000	-1.5000	14	700.0000	-17	571.5000
104.0000	103.0000	-1.0000	-11	718.0000	14	571.5000
110.0000	114.0000	4.0000	-3	724.0000	4	571.5000
134.0000	134.5000	0.5000	4	748.0000	-5	571.5000
182.0000	183.5000	1.5000	11	796.0000	-15	571.5000
194.0000	197.0000	3.0000	-4	808.0000	6	571.5000
218.0000	218.0000	0.0000	14	832.0000	-20	571.5000
224.0000	228.0000	4.0000	3	838.0000	-4	571.5000
236.0000	235.0000	-1.0000	7	850.0000	-10	571.5000
254.0000	253.5000	-0.5000	-3	868.0000	5	571.5000
266.0000	263.0000	-3.0000	-1	880.0000	2	571.5000
278.0000	278.0000	0.0000	8	892.0000	-12	571.5000
290.0000	291.0000	1.0000	-6	904.0000	10	571.5000
296.0000	294.5000	-1.5000	11	910.0000	-17	571.5000
320.0000	320.0000	0.0000	-7	934.0000	12	571.5000
332.0000	335.5000	3.5000	7	946.0000	-11	571.5000
362.0000	360.0000	-2.0000	-9	976.0000	16	571.5000
452.0000	450.0000	-2.0000	9	1066.0000	-16	571.5000
470.0000	472.0000	2.0000	-8	1084.0000	16	571.5000
476.0000	477.0000	1.0000	-9	1090.0000	18	571.5000

Figure 3. Dual conversion receiver spur table (example 1).

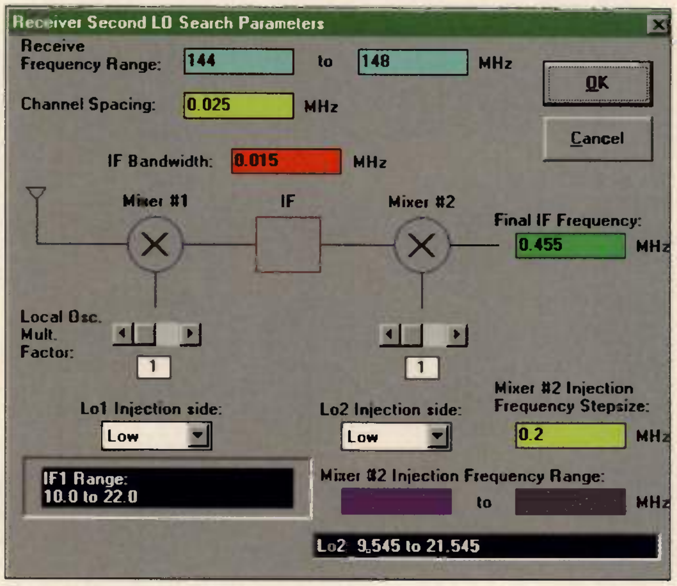


Figure 4. Search parameters (example 2).

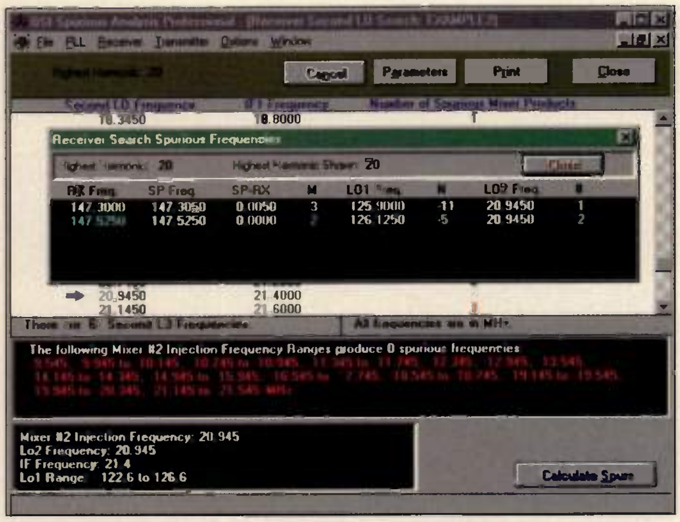


Figure 5. Search table results with "Calculate Spurs" window (Example 2).

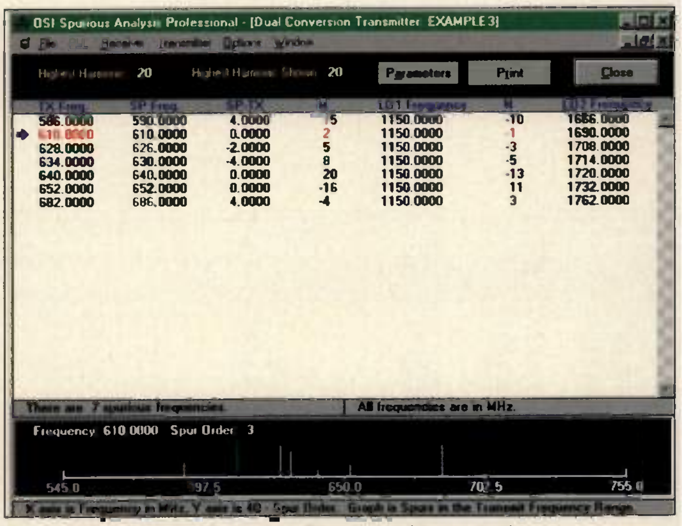


Figure 6. Dual-conversion transmitter spurs (Example 3).

the block diagram with parameters of the design reappears. Edit the appropriate values and again click on "OK." A new spurious-frequency calculation takes place with a new spectral plot. Parameter values can be saved by clicking on "File" and selecting "Save."

Example 2: search feature

The process for finding the first IF and fixed Lo in a dual-conversion design to give minimum spurs can be done automatically. Select "Search" in the Receiver or Transmitter menu and enter the values for the search. As an example, a look at a typical two-meter amateur radio VHF transceiver covering from 144 to 148 MHz presents a clear way the feature works.

Most two meter transceivers are dual-conversion with a 21.4 MHz (two spurs) or 10.7 MHz (no spurs) first IF and a 455 kHz second IF. But if another IF between 10 and 22 MHz were desired (because of an image problem or Lo leakage), new IF values may be tested using the Search option. Select "Search" in the Receiver pull-down menu and enter the values as shown in Figure 4. The results of the search for the second Lo and corresponding first IF are presented in a table as shown in Figure 5. A summary is also given that provides the ranges of the second Lo, which will produce a minimum number of spurs as set by the user. Click on any value in the table to get specific details about the design parameters, including the variable Lo range and the current fixed Lo and first IF values. For more information about the spur total, click on the "Calculate Spurs" button.

A new window will appear, showing the individual spurs associated with the selected table value and how they were calculated. For example, at an IF value of 21.4 MHz, the table indicates there are two spurs. After clicking on the "Calculate Spurs" button, two spurs are shown, one each at 147.3 and 147.5 MHz. Therefore, each spur total can be expanded upon to see where the spurs are located in the band. Because many designs will not have zero spurs but, instead, a lengthy table of total spurs over a wide range, the Search option allows for quickly narrowing in on IF and fixed Lo combinations that could provide a high quality design.

As can be seen in this two-meter design example, the search process is much quicker than continually entering new IF and Fixed Lo values and hoping to find a combination that gives the desired results. It also provides a unique way of testing general design parameter ranges, allowing one to find values that appear more promising.

Example 3: down stream generator

To design a CATV down stream generator covering the range from 550 to 750 MHz, select the Transmitter menu and click on "Dual conversion." A block diagram appears much like the one in Figure 2, except it is clearly for a transmitter with an input at 70 MHz.

The other design parameters include a channel spacing of 6 MHz, IF bandwidth of 10 MHz, first Lo injection frequency of 1,150 MHz and Lo2 injection side set at High. Fill in the values and click on "OK." As with all types of spurious calculations, the parameters in Dual conversion are error-checked to ensure proper parameters have been entered. If invalid parameters are given, the program will inform the user and suggest valid parameters. The spurious frequency table appears (Figure 6) showing the output from 546 to 754 (550 to 750 plus one-half the IF bandwidth on each end).

There are several potentially strong spurs. If one desires to see where a spur lies in the spectral plot, simply click on the spur in the table, and the spectral line associated with this spur will be highlighted in green. Or, by clicking directly on a spectral line, the frequency and order of the spur is displayed. This helps the designer to determine what portions of the band will cause the most difficulties. As shown earlier, the Search feature can be used, or individual editing of the parameter screen can be done to find a better IF and local oscillator combination.

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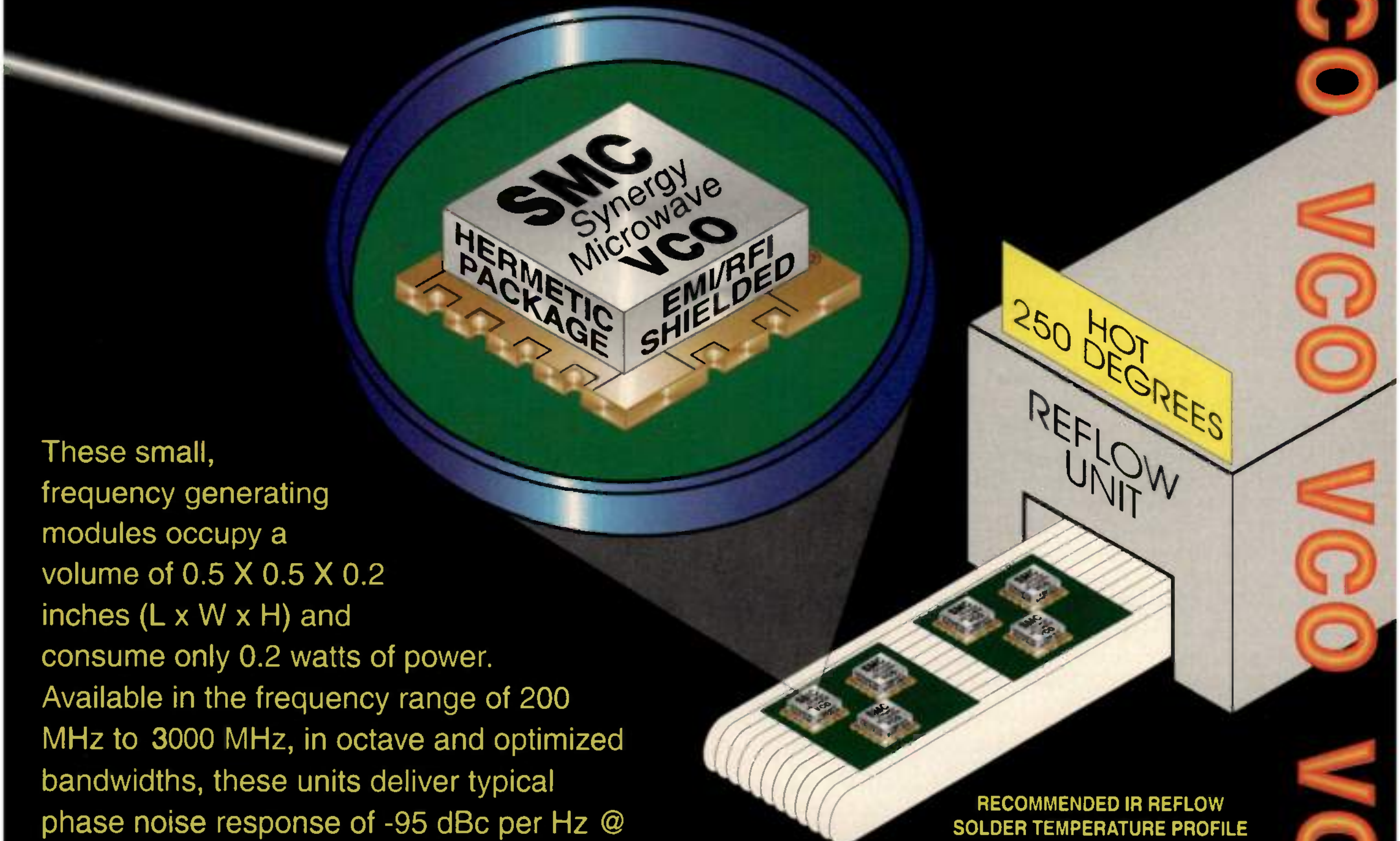

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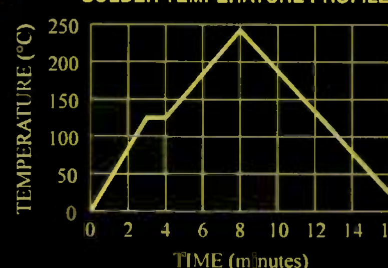

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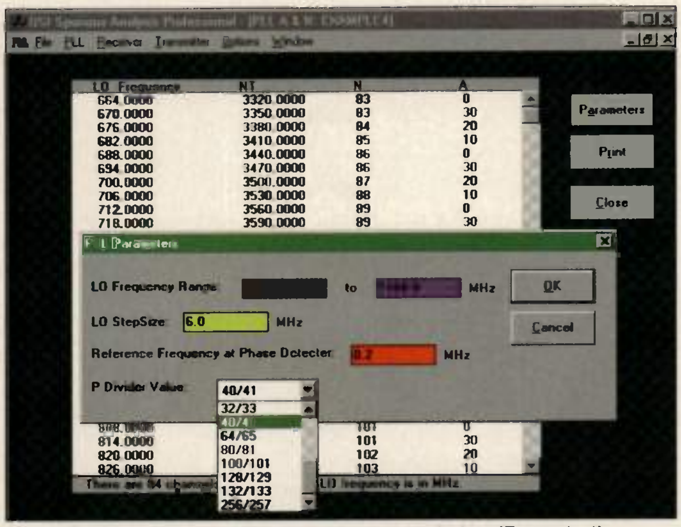

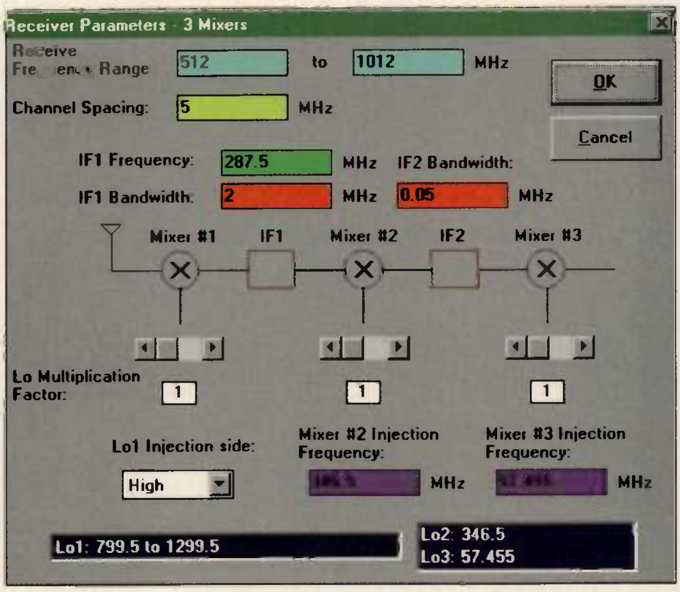

Figure 7 (above). PLL counter with parameter window (Example 4).

Figure 8 (right). Triple-conversion receiver parameters (Example 5).

Example 4: PLL counter calculations

One additional feature that is included is a phase locked loop (PLL) program to determine the selection of the best prescalar and counter combination. After a spurious analysis has been done, select "PLL" on the menu. A parameter window appears with the Lo range values already loaded in from the spurious design. Or, if the user desires, Lo values may be entered independently of the spurious pre-loaded values. All that is needed is the phase-detector reference frequency and the desired dual-modulus prescalar value. All known dual-modulus values are present, and one can be selected by scrolling to the one of interest (Figure 7). Enter the reference frequency and desired prescalar, then click "OK."

A table appears, showing the total count (NT), N and A counter values for each channel. The PLL program checks for valid parameters and results, and if the parameters or table results seem questionable, the program will notify the user. If there are no values shown in some places in the table, then the prescalar selected, reference frequency, or both needs to be changed. This feature of the spurious analysis program quickly allows the designer to find a prescalar and reference that results in values of N and A for the programmable counter to be realizable, and possibly clever. This process may result in a design that makes programming for parallel entry programmable counters or microprocessor-based serial entry more simple than first thought possible.

Example 5: triple-conversion receiver

To demonstrate the effect each parameter has on the overall design, consider a triple-conversion, wide band receiver with parameters as shown in Figure 8. Select "Triple Conversion" from the Receiver pull

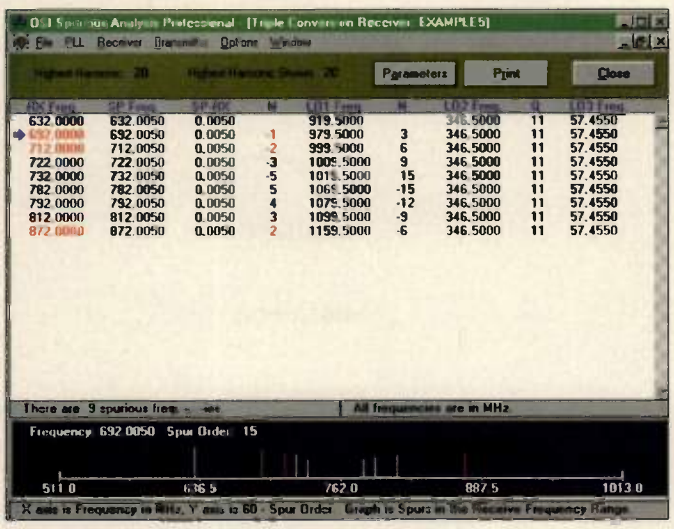

Figure 9. Triple-conversion spur table (Example 5).

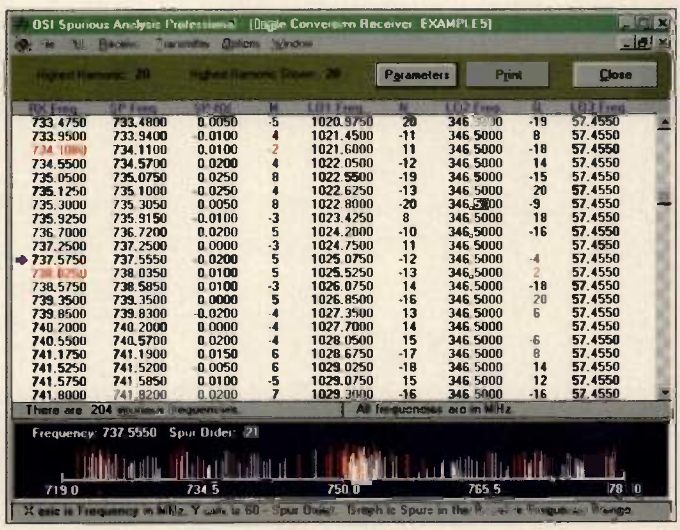

Figure 10. Triple conversion table (0.025 MHz steps, example 5).

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C3271	80-1000	50	20	1.5Kw
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down menu. After entering all values and clicking on "OK," the calculation results are shown in table and spectral line plot format as indicated in Figure 9. Initially, the parameters are set to cover a wide receive frequency range and large channel spacing to determine a general idea of what spurs, if any, the band will contain.

A wide channel spacing of 5 MHz was selected. From the table, one notices most of the spurs occur in the 700–800 MHz range, while the rest of the band appears to be clean. One can then perform the calculations using a smaller channel spacing. With the same parameters used above, except the channel spacing reduced to 25 kHz, the output table and spectral line plot change as shown in Figure 10.

It is now evident that the spurs are clumped around the spectral lines shown in Figure 9, but there are many spurs in between and throughout the entire band. Although the 5 MHz channel-spacing design appeared to be clean, simply reducing the channel spacing to 25 kHz dramatically increased the number of spurs. By changing one parameter, the characteristics of the receiver change immensely.

Again, consider the 5 MHz design. There are spurs mainly in the 700-800 MHz range. One may reduce the bandwidth or change other parameters in order to determine if the 700–800 MHz range spurs can be eliminated, or one may decide that these current spurious frequencies are of weak enough orders not to have any noticeable effects on the receiver. The designer can also use the current design for receiving frequencies 500–700 MHz and 800–1000 MHz, and use another design for the band 700–800 MHz. Several design strategies are available, and each can be scrutinized by simple adjustments to the calculation parameters.

Conclusion

Appropriate software greatly aids the designer by providing information about internally generated spurs and PLL counters. Software that takes advantage of Windows allows several different types of designs to be open and viewed at once.

RF

About the Author

Mark Atkinson received a B.S.E.E. and an M.S.E.E. from Purdue Univerity. He has worked at the Naval Air Warfare Center and is currently a software development manager at Orion Software, Indianapolis. Orion offers the software described in this article under the name Spurious Analysis Professional.

CDMA signals: A challenge for power amplifiers

By Klaus D. Tiepermann
Rohde & Schwarz

An important characteristic of power amplifiers is the efficiency of their output stages. This is particularly true for solid-state amplifiers designed for military or heavy-duty commercial use. This use imposes the conflicting requirements of low device junction temperature to insure reliability and operation at high ambient temperatures.

Inefficient operation affects the entire power amplifier package, requiring not only complex cooling schemes for the output devices, but requiring a larger power supply and, for portable applications, a larger generator as well, just to supply the extra power that is wasted as heat.

An attractive approach to minimizing the loss in the output devices is to operate them in as close to a *switch mode* as possible. [1,2]

Unfortunately, switch mode operation requires that the harmonic voltages and currents be closely controlled. If reactive (reflective) filters are used, reactive elements must be located (electrically) near the output devices. In some cases, this cannot be done. For example, in a 2-to-30 MHz, 5 kW amplifier, 16 power modules might be required, combined in a four level binary combiner. Considering cabling and combiner electrical length, there can be many electrical degrees at the harmonic frequencies between the harmonic filter assembly and the device junctions. At some frequencies, this can cause severe loss of efficiency and large voltage swings at the device junction, which can damage or destroy the devices.

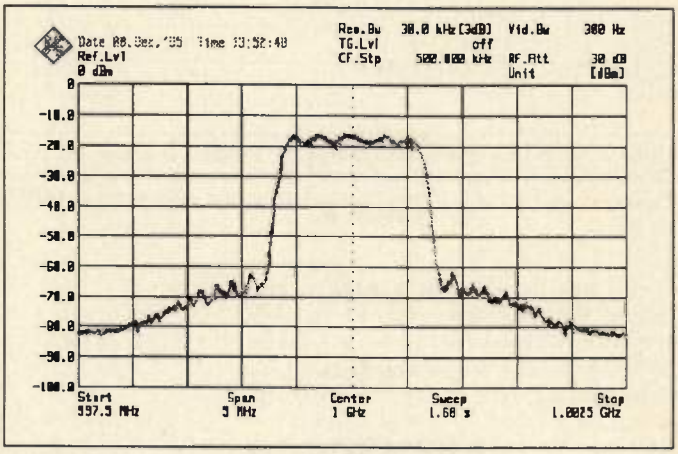

Figure 2. CDMA spectrum.

To explore this situation, we will simulate some idealized circuits to see what happens when a switched source operates in a narrowband load.

Figure 1 shows a PSPICE version of a push-pull output stage with output devices simulated by capacity-loaded switches. In a real-world circuit, of course, there would be many other circuit elements, such as DC blocking capacitors, biasing circuits and a balanced-to-unbalanced transformer (balun). In fact, this could be one of many modules driven in parallel and working into a combiner. However, the effects we wish to explore are not influenced by these elements; they only clutter up the simulation.

The switches are operated by the voltages across R_1 and R_2, V_1 and V_2. When V_1 exceeds 0.01 V, S_1 closes and looks like a low resistance. When V_1 is less than 0 V, S_1

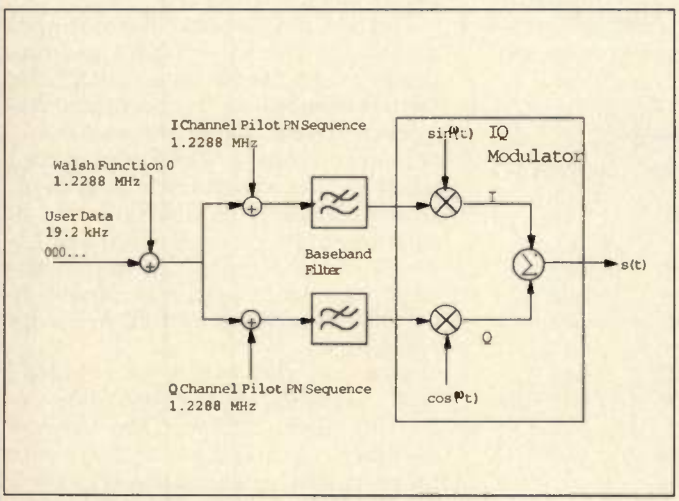

Figure 1. Generating the pilot signal of a base station.

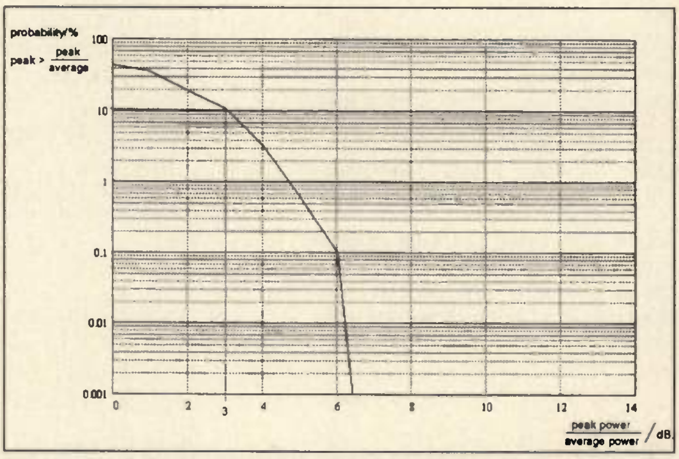

Figure 3. Statistical power distribution of a pilot signal where the probability of occurance for instantaneous envelope power values is given in decibels above the average power.

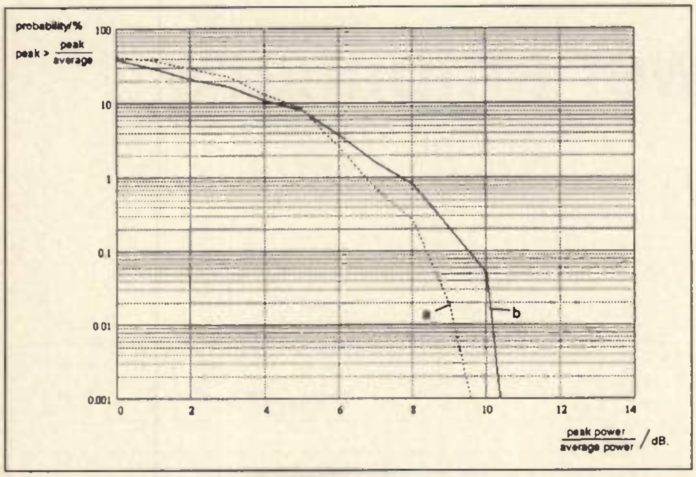

Figure 4. Statistical power distribution for a signal with two and nine CDMA channels. Curve (a) represents the pilot and traffic channels ($P_{traffic}=P_{pilot}$ − 3dB). Curve (b) represents the nine-channel CDMA signal.

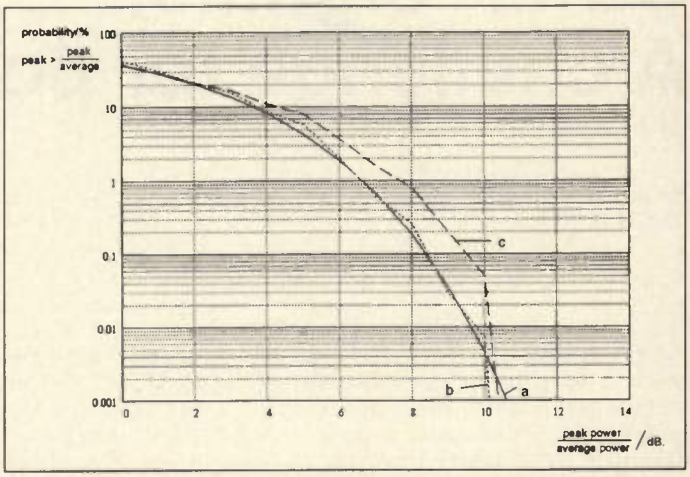

Figure 5. A comparison of statistical power distributions. Curve (a) represents the exponential distribution as theoretical distribution function of AWGN. Curve (b) represents a measured distribution of an AWGN signal. Curve (c) represents a nine-channel CDMA signal.

opens and looks like a high resistance; V_2 similarly controls S_2. We arbitrarily choose V_s to be a 1 MHz, 10 V peak sine wave and C_1 and C_2 to be 100 pF. We chose the *on* resistance, R_S, to be 1 Ω and the *off* resistance to be 1 MΩ. For reasons that will be evident later, we choose R_L = 16.667 Ω.

With L_1 and L_2 large enough to give a substantially constant current, the output across R_L is a square wave of about four times V_{DC} peak-to-peak. With V_{DC} = 28 V, PSPICE gives the spectral components and average current as follows:

I_{DC} = 5.42 A

HARMONIC NO.	AMPLITUDE (V)
Fundamental	57.50
2	0.53
3	19.20
4	0.54
5	11.50
6	0.54
7	8.20
8	0.55
9	6.30

The efficiency is the ratio of the fundamental power divided by the DC power. For this idealized circuit, the inductor current is:

$$I_L = \frac{2V_{DC}}{R_L + 4R_S} \quad (1)$$

The peak-to-peak square wave output is:

$$V_{PP} = \frac{4V_{DC}R_L}{R_L + 4R_S} \quad (2)$$

The efficiency is:

$$Eff = \frac{(0.8116)R_L}{R_L + 4R_S} \quad (3)$$

As R_S approaches 0, the efficiency approaches 81.16%. For our circuit, the efficiency is 65.3%.

We now investigate what happens when the harmonic power is not absorbed, but is instead reflected back into the switches. This is the case when a harmonic filter is added to our simple power amplifier, as shown in Figure 2.

C_3–C_5 and L_3–L_6 form a balanced version of a five-pole Chebychev lowpass filter. This one is chosen to have 0.01 dB ripple. PSPICE gives the spectral results and average current as follows:

I_{DC} = 6.208 A

HARMONIC NUMBER	ACROSS C3 (V)	ACROSS RL (V)
Fundamental	69.91	67.01
2	0.47	0.24
3	9.58	0.78
4	0.16	0.16
5	3.29	0.15
6	0.42	0.17
7	1.74	0.10
8	0.21	0.35
9	1.41	0.20

Compared to the no-filter case, the results are gratifying. The efficiency is up to 77.5%, and the harmonic power has not caused any great voltage swings on the switches. However, let us now move further into the real world. Our 2 to 30 MHz amplifier will need eight switched harmonic lowpass filters. We cannot filter each of the 16 modules before combining. Controlling the phase would be horrendous. We have to do our combining before we get to the switched filter assembly.

We simulate this situation by adding a 16.667 Ω transmission line between the switches and the filter, as shown in Figure 3.

Table 1 summarizes a lot of PSPICE running time and shows in a dramatic fashion the problems harmonic power can cause.

The line lengths transform the reactance of the filter into an inductance that resonates with the switch capacity at the harmonic frequency. The stronger the harmonic, the more voltage is generated, and the more the efficiency is degraded. Note also the dramatic reduction in fundamental power. Even if the devices could tolerate the huge voltages, the transmitter is forced to operate at as much as 4 dB below its capability.

How can this effect be avoided? One solution is to use non-reflecting filters. These are usually designed as highpass and lowpass pairs called *complimentary filters.*

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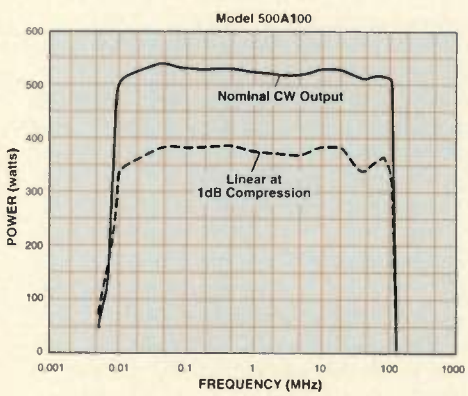

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The highpass section is terminated and serves only to absorb the harmonic energy. However, some problems are associated with the design of these filters [3]. Another consideration is the number of additional circuit components required. To maintain a low voltage standing wave ratio (VSWR), our eight filters might need nine reactive elements each. The complimentary sections might need seven elements each, adding 56 elements to our design.

The *other* approach is the subject of the following section. Consider the block diagram of Figure 4. The phase-shifting sections provide the indicated phase shift over the band F_{low} to $5 \times F_{high}$. Table 2 traces the phase shift through the diagram.

The first phase shift section shifts the reference channel (B) by a variable amount X and shifts the A and C channels by X+60 degrees and X–60 degrees as shown. We assume the signal is sinusoidal at this point in the amplifier, so the phase shift of the harmonics is of no interest. We make the assumption that the power amplifiers (PAs) generate square waves in phase with their inputs. The phases shown are with respect to the reference channel phases. Note that in any channel, the shift of the Nth harmonic is N times the shift of the fundamental.

Each channel goes through a last phase shifter as shown. Here, the reference channel is shifted by a variable angle Y, and the other two channels are shifted by Y–60 degrees and Y+60 degrees.

The three channels are summed in the output combiner, with the fundamental and seventh harmonic power adding in the output load and the other harmonic power being dissipated in the combiner dump ports.

This scheme substantially does what we want. The only potential problem is the seventh harmonic, but it is not strong to begin with, and in a real transmitter, insertion loss and filter Qs minimize its buildup.

To test this scheme, a PSPICE file was generated for the circuit shown in Figure 5. In this circuit, three push-pull switched sources drive three networks, which shift the phase as shown. Their outputs are combined in a tri-hybrid. The switches are operated by a 3 MHz

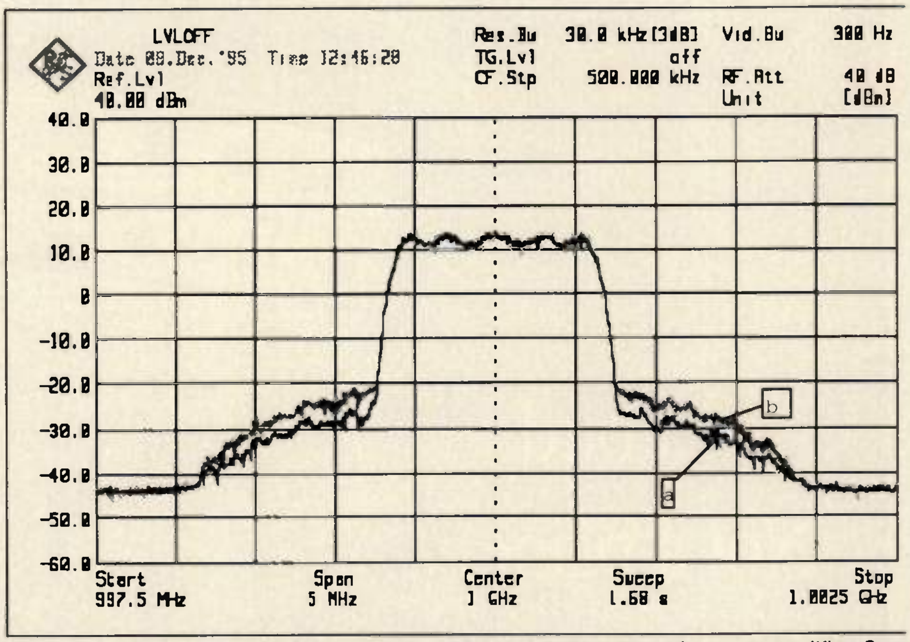

Figure 6. CDMA spectrum with intermodulation sidebands at the output of a power amplifier. Curve (a) represents a pilot signal with an output power of 1 W. Curve (b) represents pilot and traffic channels (–3 dB) with a total power of 1 W.

source with outputs at 0°, –60° and –120°. The simulation used delay lines to get the phase shift, because only one frequency was being run. In a real amplifier, constant phase-shifting sections would be used. To minimize running time, only a single two-pole circuit was used in each channel. This limited the bandwidth ratio but still allowed fair accuracy up to the ninth harmonic.

The test was intended to answer the following questions:

a) What is the harmonic level at the switches? Is the square wave well-preserved?

b) What is the harmonic level at the output? Is it sufficiently low so that reflected harmonic power is of no consequence?

c) What is the conversion efficiency, DC to fundamental power?

Table 3 is taken from the PSPICE printout. V(7,8), V(9,10) and V(11,12) are the spectral components at the switches and show fairly well behaved harmonics.

V(2,0) is the output showing a fundamental of 38 V peak with minuscule third and fifth harmonic and a seventh harmonic more than 16 dB down.

V(40,41) shows a 3.58 A DC current, giving an input DC power of 5(3.58) = 17.9 W. The fundamental power is 38(38)/2/50 = 14.44 W giving 80.6% efficiency.

We can conclude that the scheme works. If we can design networks that provide relative phase of –60°, 0° and 60° over a broadband, we can implement the blocks of Figure 4 and cancel, or at least considerably reduce, third and fifth harmonics.

How do we design the networks? There may be an analytic way to synthesize them, but I'm not aware of it. One way that works reasonably well is to find the pole frequencies for a two-channel, 120° network, then add in the reference channel using approximate geometric mean pole frequencies. An example will make this clear. Assume we want to build a 2-to-30 MHz amplifier. The output constant-phase allpass networks will then have to cover 2 to 150 MHz to pass the fifth harmonic.

Using an allpass network program such as *RF Design's* RFD–0193, we find that 4 poles/channel will give us 120° with less than 0.2° error over the band. The pole frequencies are:

We find the geometric mean of each pair as trial pole frequencies for channel B. This gives us:

CHANNEL B (MHz)
251.0060600
36.0400560
8.3240690
1.1951886

The phase shift produced in each channel is:

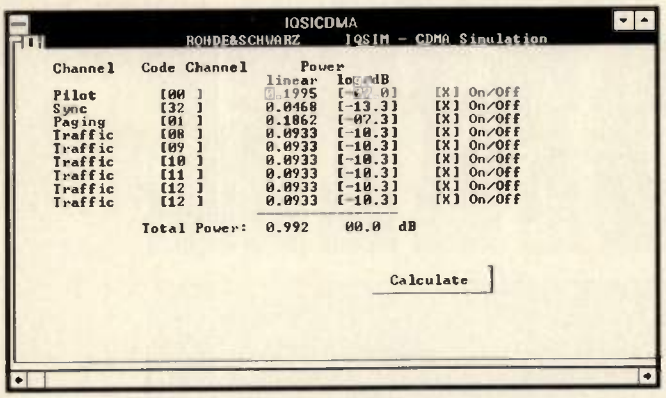

Figure 7. Configuration menu for multichannel CDMA signal.

$$\arctan\left(\frac{P_1}{F}\right)+\arctan\left(\frac{P_2}{F}\right)+\arctan\left(\frac{P_3}{F}\right)+\arctan\left(\frac{P_4}{F}\right) \quad (4)$$

where P_1–P_4 are the pole frequencies in that channel, and F is the frequency.

Subtracting the shift in channel A from the shift in channel B, we find that the difference ranges from 55°–65° over the 2-to-150 MHz band. In an effort to reduce the error, we slightly shift the pole frequencies of channel B. We find that reducing P_1 and P_2 by 0.933 and increasing P_3 and P_4 by the same factor gives an error of less than 1° over the band. This is certainly accurate enough for our purposes. With the pole frequencies established, we can easily obtain the circuit values for our choice of network.

Conclusion

A scheme has been presented that enables switch-mode operation of output devices over a broad bandwidth without tuning. A method has been shown for designing the critical phase shifting circuits. *RF*

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About the Author

Klaus D. Tiepermann is with Rohde & Schwarz. The company makes the SMHU 58 signal generator, the ADS arbitrary waveform generator, and IQSIM-K software described in this article. Alternatively, a Tektronix AWG 2021 can be used as the ARB generator.

Coping with cost constraints in a market-driven economy

By Ernest Worthman
Contributing Editor

In today's market-driven economy, the design engineer quickly finds out that *cost* is usually the most significant issue that must be faced.

In engineering school, we spent many a night trying to figure out the Fourier transform of a step-pulse function for a digital switch. But until recently, few of us ever sat through even a *fundamental* economics class.

So, once we hit the real world and proudly design the voltage-controlled oscillator (VCO) with the flat power curve across its full bandwidth, or the near-perfect linear microwave power amplifier, the order comes down to cut 30% out of the production cost. *Baptism by marketing.*

Marketing perspective

In most businesses, marketing is the guiding force. Marketing wants your product to cost nothing, do everything, be 100% user-friendly and be 100% reliable.

We all are aware of the absurdity of such a concept. It is, however, marketing's job to come as close to this equation as possible. So you, the engineer, are under tremendous pressure to make the product as competitive as possible. If you can't design it competitively, they can't sell it, and your company makes no money.

Because engineers are, by default, mathematically perspicacious, taking the project costs and plugging them into an equation is a good way to analyze the project. Some costs are beyond the engineer's control, but your marketing department will be impressed by a well-developed table of costs.

One fairly generic equation serves as a good example. Of course, the equation can be adjusted to include situation-specific requirements.

First, we have to decide which factors to include. Once the factors are determined, we can assign variables to them and work them into the equation.

On the macro scale, take a look at the global picture. All organizations have fixed costs, i.e., costs that are always present. Items such as power, lease costs, building maintenance, insurance and certain labor costs are typical examples. These costs can be lumped and represented by variables. These costs usually are not under the direct control of the engineer; instead they become fixed variables in the equation.

Second, let's address the costs that usually *are* under the control of the engineer. I call them *local costs*. Examples are materials, design costs, parts warehousing, quality control, test-and-analysis, and support costs.

Finally, there are other variables such as profit, environmental costs, international delivery costs (if applicable) and regulatory costs (such as FAA, FCC and EPA) that may need to be considered.

Once the costs are identified, we can develop an equation to help us to analyze the project. The fundamental equation is :

$$C = P(G[L + O + M]).$$

where

L = Labor
O = Overhead @100% of Labor
M = Materials
G = G&A (general & administrative)
P = Profit
C = Selling price

The next step is to develop the equation. From there we can add other variables as needed. Let's assume that there are no padding costs or miscellaneous costs—yet. So the first thing we will do is to add a 10% G&A (G = 1.1). Next, we want to add a 25% profit (P = 1.25). Now, to give some substance to this, let L = \$1, O = \$2 and M = \$1.50. If we plug these into the equation, we come up with a selling price of \$6.19.

If it becomes necessary to add other costs to the project, include them based on their relationship to the basic production cost—that is, make sure the variable isn't in the wrong place in the equation. Otherwise, the equation may not reflect the actual cost.

As engineers, let's take a look at the cost considerations we have the ability to affect.

Engineer's perspective

From an engineering perspective, some of the earlier-mentioned variables can make the actual cost determination a bit more difficult. This is because the engineer sometimes is not given all of the information. Therefore, we need to analyze which additional factors must be included and where.

The product being designed is a Whutzit. Design specification call for a combined RTC and a linear RF amplifier for 2.4 GHz wireless network applications. You must bring the product to market in six months and for less than your competition's cost of \$6.50.

You analyze the requirements, and you are given the fixed costs. For the sake of discussion, the earlier figures will be used.

One of the additional variables you must consider is *parts warehousing*. Because your facility is small, you must buy parts in smaller quantities and at slightly higher cost. This adds 5% to the material cost (M = 1.05*1.5 = 1.575). Warehousing cost is a factor of the material costs and, therefore, affects only the material costs. Now the selling price is \$6.27.

The device also requires FCC approval. The delay and additional labor to prepare the FCC submission add 3% to the G&A (G = 1.10 + 0.03 = 1.13). This brings the selling price to \$6.45. Notice that this G&A expense has a profound effect on the final cost, even though it is a small percentage.

Now, you think that you have determined all of the costs and come up with a final figure. Just as you are getting ready to present your picture to management, your are informed that there will be a strike in an overseas plant that delivers semiconductor substrate. You are informed that there will be an eight-week delay before the first shipment of substrate arrives. Your choices are to halt production for eight weeks or to purchase substrate elsewhere at a 100% premium above your current supplier's price. You can either absorb

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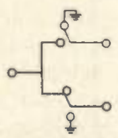
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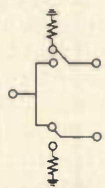
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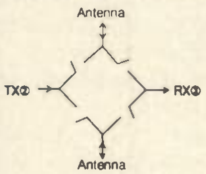

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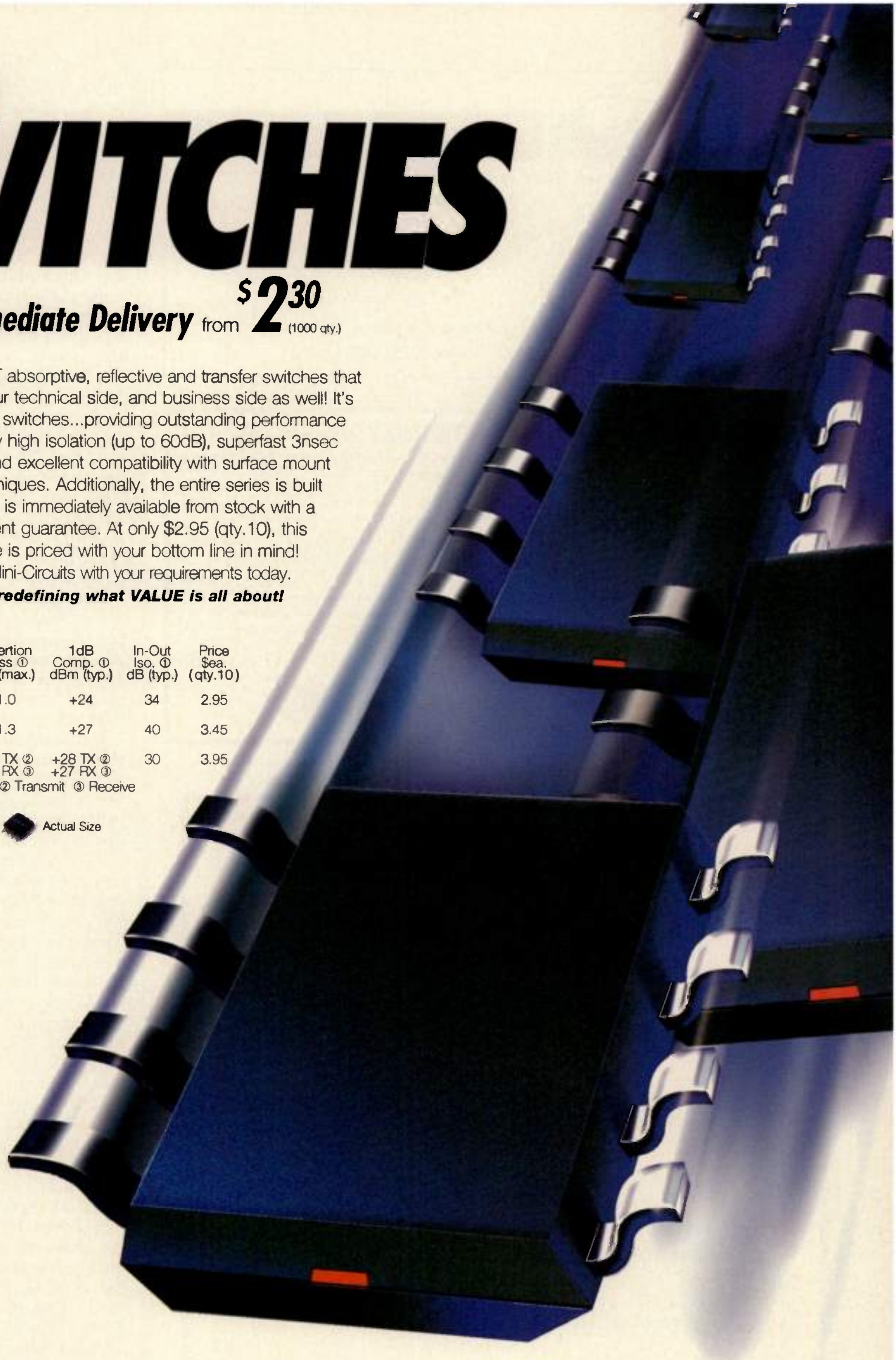

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the downtime or purchase more expensive material to keep the labor busy and the lines moving, but this will add cost to material. (Substrate is 50% of material costs.)

Leaving profit out of the equation, let's analyze both of these options and see which one is more cost-effective. Let the equation help you to determine the direction. Your company can produce 4 million units during the first year, or 333,333 units per month. One option is to go ahead with production for the strike period using the more expensive substrate. In this case, modify the basic equation by adding the additional material cost (M = 1.57 + 0.78 = 2.35). This yields a final production cost of $6.04 (vs. $5.16). Note that this is a 15% cost increase, 88 cents higher than the original cost. It also puts the new selling price (1.25*$6.04 = $7.55) way above the competition's. This will be your first two months' cost if you decide to go ahead with production. To get a total production cost, you add the first two months' production cost to the other 10 months' production cost. The yearly production cost is then [2*(333,333*$6.04)] + [10*(333,333*$5.16)] = $21.21 million. Taking a quick look at the original cost, without the strike, it was [12*(333,333*$5.16)] = $20.6 million. Finally, doing the overall math, it works out to about $610,000 more, or about 15 cents per unit.

Next, go through the above process again, but work the equation with the option to delay the production (M = 0). This will have a two-month delay upon the product. This will cost the company $3.39 (C = [1.13(1+2+0)]) for each unit that cannot be produced because the company still has to cover the fixed costs and support equipment and staff. Therefore, the cost of delay is: 2*(333,333*$3.39) = $2.26 million. Next, add the cost of the original 12-months' production cost of $5.16 per unit and you get: 12*(333,333*$5.16) + 2($3.39*333,333) = $22.26 million in production costs, which is $1.66 million more in production costs and about 41 cents per unit.

Because it is pretty obvious that the more cost-effective way is to go with the second substrate source, we can come up with a final selling price: C = [1.25(1.13(1+2+1.57)] + (0.15) = $6.60 (This time the additional cost is added as a stand-alone variable in the equation because it was amortized over the entire production run as a fixed cost.) This result is 10 cents above the competition's price. This result is what we take to management.

At the management meeting, you present your analysis. You include the obligatory charts and supporting documents showing how you arrived at the new cost. Marketing insists that the selling price cannot be over $6.50. You stand your ground, in light of the fact that you need some additional resources to make the unit, in light of the unforeseen circumstances.

Compromise

You try to work a compromise, and after much debate, marketing agrees to give up 3 cents per unit. That leaves you to find a way to cut the other 7 cents. This is when you look at your design costs, you work with suppliers and you try to lower some of the variable in your equation.

What to look at — Many areas can offer cost savings. Two of my favorites are *design costs* and *parts costs*. Design costs include labor and materials (other than assembly parts). Although it is not popular to do so, look is at the design team. Can the design team be made more efficient? Can employees be reassigned to another project, or can their cost be shared with another product, without affecting the time line of the development cycle? Or, can some tools (such as software) be purchased up front, tools that will increase the efficiency of the design team? And, can an existing design be adapted, as well as existing fixtures for manufacturing? If such options exist, they can be plugged into the equation in the proper place, and you can see the effect on the final cost.

Now, let's say that the percentage makeup of the design team part of the labor (L) looks like this:

STAFF	
3 engineers	45%
1 CAD draftsman	15%
1 support staff	10%
1 manager	3%
Total	73%

This means that the design team makes up 73% of the total labor (L). Through some redesign efforts you take 10% off of the engineers' costs. Therefore, 10% of 45% = 4.5%. This makes the labor for engineering cost 40.5%. Your staff cost is now 68.5%. If labor were $1 before, your share was 73 cents. Now it is 68.5 cents. That takes 4.5% off of the $1 to make it 95.5 cents. When we plug this into the equation, we get $5.12 and a selling price of $6.40. As it turns out, this took 4 cents out of the original production cost of $5.16. That leaves another 3 cents in cost reduction to find. Adding the 15-cent amortized substrate cost brings the selling price to $6.55. Backing out the 3 cents from marketing brings it down to $6.52. (The 1-cent discrepancy is due to rounding off the figures).

Now you only have to find another 2 cents in other areas. Of course, you can continue to look at labor, but for the sake of argument, assume that this is all you can find to cut from it.

Another area ripe for cost cutting is *parts*. There are a number of ways to cut costs, and the most obvious is to try to get cheaper components or to reduce the component list. Let's assume that we have already done that, and any further compromise in component integrity will put us outside the acceptable performance delta.

If we can't reduce component cost, maybe we can reduce component handling. As mentioned before, the facility is too small to warehouse a lot of parts and, therefore, we are paying a slight premium to order smaller quantities. Perhaps we can institute a JIT (just-in-time) scheme that will allow us to place a larger order and get a better price break.

After discussing this with the various vendors, it turns out that we are able to use a JIT program with some of them. In doing so, we get a 2.5% overall concession in parts prices. That drops the total parts cost from $1.57 to $1.54 ($P = 1.54$). Again, plugging these numbers into the equation gives us $5.09. Notice that this took out an additional 3 cents, which is enough to bring the production cost under the line. Figuring in the profit gives us a selling price of $6.35. Finally, again adding the 15-cent amortized substrate cost, the selling price is $6.50, which is the target price.

Conclusion

An infinite number of variables can be used to replace or complement the examples and to modify the equations. There are also other equations that can be applied under the same concept to achieve similar results.

Engineering no longer is a *design first, cost later* profession. Today's engineers must be extremely cost-conscious. Novice engineers must concern themselves not only with design, but with cost-effective design.

In preparing this article, I used information supplied by Andy Przedpelski of the Shedd Group and H. Clark Bell of hf+, an RF and microwave engineering and consulting firm in La Jolla, CA. RF

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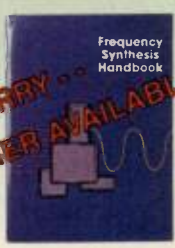

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Production testing grows as an RF product's design requirement

By Gary A. Breed

Design for manufacturing always has been part of RF engineering, but not always a big part. Many traditional RF products have been designed almost entirely for performance and have been manufactured in relatively small quantities. Radio and TV transmitting equipment, test-and-measurement instruments and military communications systems are good examples.

With changes in the market for radio-based wireless communications services, most RF manufacturing involves commercial or consumer products. Designers must include "manufacturability" as a key design parameter. RF products typically require more testing, and testing with more types of equipment, compared to other electronic devices. To reduce costs, testing must be simplified by simultaneously reducing the number of tests and increasing the speed of the rest.

No-tune circuits

Part of RF production testing consists of observing measurements while making adjustments. This manual procedure is inefficient and expensive. Several design techniques can be used to simplify or to eliminate adjustments on the assembly line.

Broadband circuits—Tuning usually means adjusting an inductor or capacitor to achieve resonance or to optimize coupling in a tuned circuit, filter or matching network. During the past 10 years, much attention has been given to broadband devices. Darlington amplifier blocks with 50Ω input and output impedances over hundreds or even thousands of megahertz have replaced discrete transistor stages in many applications. Despite increased current consumption, these devices can reduce component count and eliminate tuning. Broadband design techniques can be applied to matching networks, even those with topologies (and component counts) the same as narrowband networks. In a broadband network, component values can vary without significantly affecting circuit operation.

Pre-tuned components—Previously, few RF products were made in quantities sufficient to warrant investing in process development, precision manufacturing equipment and automated testing. Now, filters, couplers and active circuit-building blocks are offered as drop-in components. The concentration of activity in the cellular band and nearby mobile radio and Part 15 Industrial Scientific and Medical (ISM) bands have created families of "commodity" components. Buying a transmit-and-receive diplexer or a Voltage-controlled Oscillator (VCO) is as easy as buying a resistor.

High-level integration—"Radio-on-a-chip" is a goal of integrated circuit-makers that guarantees a no-tune situation. The challenges of putting RF on a monolithic silicon or GaAs wafer are significant: combatting high-frequency coupling effects, the inability to make large capacitors for coupling and decoupling, the size and low Q of monolithic inductors, and a lack of RF experience among chip-makers that for a long time have served only digital and low-frequency analog markets. Still, available devices include virtually all RF circuitry on-chip, with only a few capacitors and a filter or two external to the device.

Computer-aided design

CAD software has two benefits for RF design. First, software circuit models are close to actual, as-built circuits. This accuracy allows a design to be analyzed and optimized for assembly using components with an expected spread of values. Designs then become more predictable. Assembly line adjustments that formerly were required "just in case" are unnecessary.

The second benefit of good CAD tools is *engineering productivity*. When software accurately represents real circuits, few prototype iterations are needed. Some of the time saved can be applied to further circuit refinement. More design options can be explored; and techniques less familiar to an engineer can be investigated, resulting in a final design closer to ideal than would be realized if the engineer were required to "stop designing and start building" at an earlier stage.

The other side of the production testing issue concerns the speed and simplicity of the tests that remain after creating the final design.

Instrumentation advances

Automated test systems—Whether done by interconnecting individual instruments or by using a modular system like VIXbus, automation speeds measurements. Computer control of instrumentation and of data analysis is both fast and simple, compared with manual operations. It reduces the skill required of the operator from that of a specialized test engineer to that of a good assembly-line worker. Virtually all high-quality instruments are capable of being used in this way.

Instrument speed—To test even faster than automation allows, individual instruments must operate faster, with faster frequency sweeps, faster changes between measurement setups, and faster settling time per measurement point. In power measurement, testing can be accelerated by using diode detectors instead of thermocouple detectors. In spectrum or network measurements, sweep time is reduced by varying the sweep rate or accuracy across the measurement band. An example would be to take measurements only in the center of the passband or at the band edge.

Multifunction instruments—As have component makers, instrument companies have created standard test sets for widely-used applications, such as cellular. Coordinating testing in one box is much easier than controlling several instruments and, in most cases, it is much less expensive.

Conclusion

High-volume manufacturing of RF products has brought changes to production that must be addressed. Designers must provide part of the cost savings in production by designing products that need a minimum of testing. This step alone isn't enough to meet production testing speed requirements, so instruments themselves are changing to meet the need of manufacturers serving a growing RF marketplace. RF

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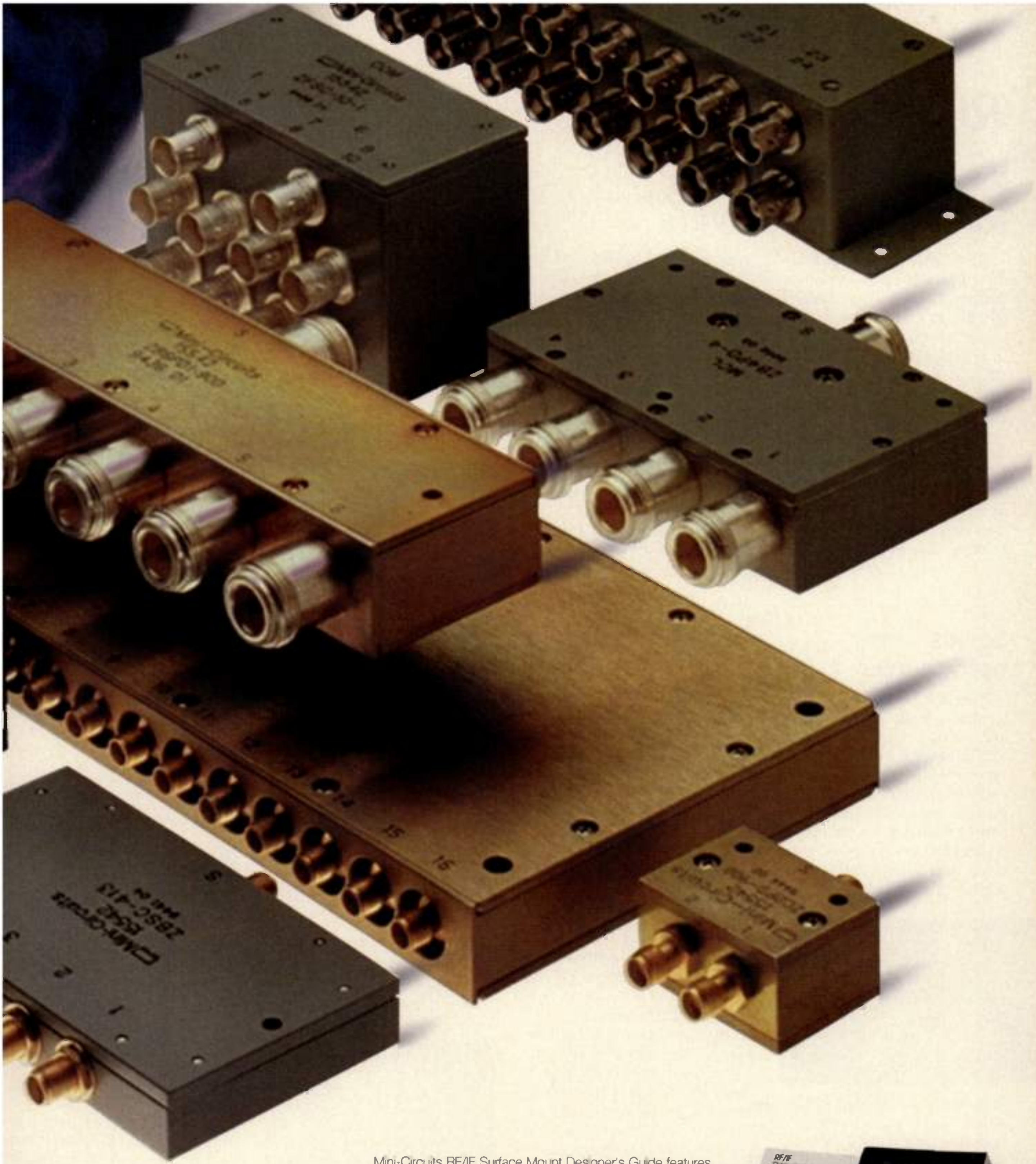



RF/IF Designer's Handbook

Front-end SAW filters for wireless receivers

RF Monolithics has introduced a series of low-loss, surface-acoustic-wave (SAW) RF filters that solve out-of-band interference problems in low-power radio systems. RF1171 and RF1172 at 418 and 433.92 MHz, respectively, have an insertion loss of 3 dB. RF1210 and RF1211 at 303.825 and 315 MHz, respectively, have an insertion loss of 1.7 dB. RF1181 at 916.5 MHz has an insertion loss of 4 dB. All of the filters use selective null placement to provide greater than 40 dB suppression of the local oscillator and image responses in superheterodyne receivers operating with a 10.7 MHz IF. The filters also have a minimum 3 dB bandwidth of 500 kHz and an ultimate rejection of 80 dB. The filters are designed for wireless remote control, meter-reading, automotive keyless entry, security, and data telemetry.

RF Monolithics
INFO/CARD #153

Series linear optocoupler

The LOC series linear optocoupler made by CP Clare's semiconductor group has applications that include instrumentation and process control (isolated 4–20 mA control loops), isolated power supply feedback, motor speed control and audio signal coupling. Coupling analog and digital signals, the LOC also

features a >200 kHz bandwidth, high gain stability, and low input and output capacitance. The device is designed for isolation amplifier designs as a transformer replacement for telephone line coupling, power supply feedback and audio signal interfacing. The LOC operates from –40° C to 85° C and is priced at $1.29 in 10,000-piece quantities.

CP Clare
INFO/CARD #154

Jitter generator

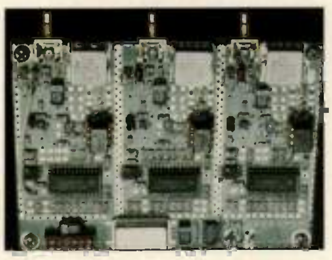
The JitterGEN jitter generator from Noise Com is an automated instrument that adds controlled jitter to data

signals at T1/E1 frequencies. The instrument stresses the T1/E1 connections to PSTN, cellular and PCS base stations, satellite modems, and microwave links. It is used for testing according to the full specifications of AT&T Accunet TR62411 and other international specifications such as G.823 and G.702. The JitterGEN provides controlled FM and PM jitter in a range from 0.009 Hz to 400kHz, and unit interval (UI) controlled jitter from 0 to 200 UI. The system will function as a stand-alone instrument or under the control of an external instrument controller. Functions are software-driven.

Noise Com
INFO/CARD #155

Miniature multiple synthesizer

RF Prototype Systems' small, low-cost multisynthesizer allows implementation to double- or triple-conversion radios and up or down converters. It features three independent PLLs on one board, each settable to a frequency from 100 and 2,800 MHz via jumpers. Each of the three RF outputs are capable of delivering 5 dBm. Isolation between each PLL is >60 dB. There is an on-board reference oscillator but the synthesizer will accept an external reference oscilla-

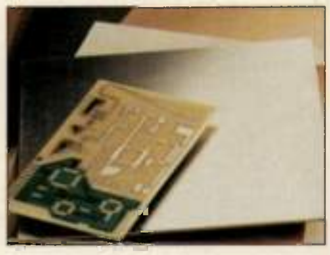


tor as well. The unit requires only one 5 VDC power supply. It is available as a board or module, and it has a reference frequency range of 1 to 20 MHz.

RF Prototype Systems
INFO/CARD #156

High-frequency laminate

RO4350, a new high-frequency circuit material



laminate from Rogers, offers high-frequency performance—compared to woven-glass PTFE substrates —with the ease of fabrication associated with epoxy and glass laminates. This material family is specifically targeted to low-cost, high-volume, commercial wireless applications. RO4350 does not require special through-hole treatments or handling procedures, which allows fabrication and assembly of high-frequency circuits for about the same cost as epoxy and glass. RO4350 features a K' of 3.48 and a thermal coefficient of expansion in the Z-axis of 50 ppm.

Rogers
INFO/CARD #157

SIGNAL PROCESSING COMPONENTS

Dielectric resonator quadroplexer

K&L model 6DRZ-725/1275P is a highly selective dielectric resonator quadroplexer with extremely low loss. Maximum insertion loss is less than 2.2 dB at each center frequency. The 3.0 dB bandwidth is 55 MHz, VSWR is 1.5:1 typical, and stopbands are 60.0 dB at Fo ±80 MHz. The unit measures 2.0" × 1.7 " × 0.8".
K&L Microwave
INFO/CARD #158

Power splitter, combiner

Mini-Circuits has introduced the PQW-2-270, a new two-way 90° power splitter and combiner featuring a wide 3:1 bandwidth covering the 90 to 270 MHz frequency range. This 50Ω device has 1.0° typical phase unbalance, 1.4 dB amplitude unbalance and 1.25:1 VSWR. The plug-in unit is housed in a hermetically sealed case and withstands operating and storage temperatures from –55° C to 100° C. Uses include I&Q and QPSK modulators. The unit is priced at $29.95 each.
Mini-Circuits
INFO/CARD #159

Multi-octave stripline power divider

Midisco model MDC2245A is a two-way isolated power divider and combiner that covers 0.5 to 4.0 GHz in one unit. The isolation is 20 dB minimum; insertion loss is 0.5 dB maximum; amplitude balance is 0.2 dB maximum; and the phase balance is 4°. VSWR is 1.3:1 maximum at the input and 1.2:1 maximum at the output. With a 1.2:1 VSWR at the load, the average power rating is 30 W. The price is $249 each.
Midisco
INFO/CARD #160

Directional couplers for CDMA feed-forward amplifiers

Merrimac has developed directional couplers specifically for feed-forward amplifier applications in CDMA systems that require a ± 0.1 dB frequency sensitivity across the operating band. The family of couplers includes coupling values to 40 dB and covers either the 900 MHz or the 1,800 MHz PCS frequency band. All have been developed for high-volume, low-cost pricing and for reliability.
Merrimac Industries
INFO/CARD #161

Surface-mount level detector

Model LNJ9901 threshold detector from Amplifonix accurately measures RF signal levels from 20 to 2,000 MHz with a sensitivity of 0.9 V. It provides a video output proportional to the input power, which is useful for signal-monitoring applications including amplitude detection, detector log video amplification and automatic level control. A schottky diode detector, voltage reference and temperature-compensated comparator are contained in a hermetically sealed, surface-mount package. The small-signal VSWR is 1.8:1. The unit operates at 15 V. The unit is priced at $111 for 1–24 units.
Amplifonix
INFO/CARD #162

CABLES & CONNECTORS

Superflexible cable

HST4-50, a 1/2" high-power superflexible cable, fits wireless applications. Made by Andrew, the HST4-50 cable

allows operation at an average power of 2.7 kW at 960 MHz. The cable features a helically corrugated, solid-copper outer conductor for better flexibility, shielding effectiveness and impact resistance when compared to braided cable. Because adhesive is not used on the inner conductor, connector installation times are decreased, and attenuation is improved.
Andrew
INFO/CARD #163

HN series RF connectors

HN series RF connectors for applications ranging from semiconductor wafer processing systems to instrumentation, cellular and wireless communications equipment are available from Tru-Connector. These connectors are slightly larger than standard N types. They handle up to 5,000 V, provide 50Ω impedance and operate over a 0-to-4 GHz frequency range. Designed to fit cables from 0.195" to 0.945" O.D., they are available as straight-through and right-angle plugs, jacks and plug receptacles, and as in-series and between-series adapters.
Tru-Connector
INFO/CARD #164

Field-replaceable SMA connectors

Coaxicom makes panel-mounted (flange), field-replaceable SMA connectors in a wide variety of configurations. These SMA-series types are made of stainless steel, in accordance with the requirements of MIL-C-83517- and MIL-C-39012. The variety includes male and female versions, with flanges ranging from 0.375" square and 0.500" square, to 0.375" × 0.500" rectangular and the two-hole flange 0.223" × 0.625". Field-replaceable TNC and N types are also available.
Coaxial Components
INFO/CARD #165

SEMICONDUCTORS

AGC amplifiers

Qualcomm offers two automatic gain control (AGC) amplifiers designed specifically for dual-mode code-division, multiple-access, frequency modulation (CDMA-FM) cellular applications. The Q5500 (receive) and Q5505 (transmit) AGC amplifiers are designed to maintain consistent power levels over changing distances between subscriber handsets and base stations and in the

presence of interference caused by multipath fading, terrain topology and environmental conditions in a cellular network. Product specifications are compatible with the IS-95 standard for CDMA digital cellular communications. They are priced at $2.95 per device in 1,000-piece quantities.
Qualcomm
INFO/CARD #166

Low-voltage voice CODEC

The ST5090 voice coder-decoder (CODEC), manufactured by SGS-Thomson, is designed for digital cellular and cordless telephone applications. The device operates on 3.3 V supplies and consumes 21 mW in operation while drawing only 0.5 μa at 3.3 V in standby. Inside the IC is a complete CODEC and filter system, including 14-bit linear A/D and D/A converters, 8-bit μ-law or A-law companding A/D and D/A converters, transmit and receive bandpass filters and an active anti-alias noise filter. Additionally, the chip includes all other audio front-end functions needed in telephone terminal applications, including microphone selection switch and preamplifier, earpiece driver, sidetone circuit, turn on-off transient suppressor, ring-tone generator and buzzer output. The unit is priced at $4.96 for SO-28 and $5.46 for TQFP 44 in 10,000-piece quantities.
SGS-Thomson Microelectronics
INFO/CARD #167

Analog multiplexers

The Maxim MAX4518 (4-channel) and MAX4519 (dual 2-channel) precision analog multiplexer units offer 100Ω maximum on-resistance, fast switching (transition times < 250ns) and ESD protection in excess of 2,000 V. They operate with a single supply of 2.7 to 15 V or with dual supplies of ±2.7 to ±8 V. Prices start at $1.15 in 1,000-piece quantities.
Maxim Integrated Products
INFO/CARD #168

Current-feedback operational amplifier

The Comlinear CLC446 current-feedback operational amplifier operates at 50mW quiescent power, offers 400 MHz bandwidth, a slew rate of 2,000 V/μs and 900 ps rise and fall times. The unit is ideal for NTSC and PAL video switchers and routers and is priced at $1.95 in 1000-piece quantities.
Comlinear
INFO/CARD #169

AMPLIFIERS

Cellular multichannel feed-forward amplifier

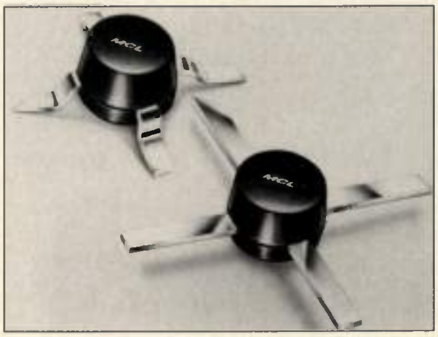
MPD Technologies linear multichannel amplifiers for digital and analog cellular applications use feed-forward techniques to achieve ultra-linearity. The solid-state amplifiers operate over the 869–894 MHz band at 35 W average output power. The amplifiers also feature 60 dB gain, a DC input of 22–27 VDC at 20 A, an IMD of 60 dBc and a size of 5.2 " × 13.7" × 16.9".
MPD Technologies
INFO/CARD #170

High-power linear amplifier

A class A linear amplifier from Comtech minimizes spectral regrowth in digitally modulated communications networks. The ARD957128-200 operates over the 950 to 1,425 MHz frequency band with linear output power of 200 W. Typical intermodulation distortion products are –28 dBc. An integral RS-422 interface for both control and remote monitoring is included.
Comtech Microwave Products
INFO/CARD #171

Monolithic amplifiers

Mini-Circuits offers two monolithic amplifiers in the DC-to-3,000 MHz frequency range. The ERA-3 is a drop-in unit typically delivering 20.2 dB gain with 1.7:1 VSWR at the input and 1.8:1 VSWR at the output. The ERA-3SM is a surface-mount design with typically 19.4 dB gain and 1.8:1 VSWR at the input and 1.9:1 VSWR at the output. Their wide bandwidth eliminates the need for costly compensation networks and extra gain stages. The units are priced at $1.67 each for the ERA-3 and $1.72 each for the ERA-3SM in quantities of 30.



Mini-Circuits
INFO-CARD #172

Linear RF amplifiers

Six linear hybrid amplifiers from Motorola include models MHL8015, MHL8018, MHL8115 and MHL8118, designed for linear applications in 50Ω systems requiring wide bandwidth, low noise and low distortion. The other two, models MHL9125 and MHL9128, were designed for high-performance, cellular base-station applications. Bandwidth ranges are 40–1,000 MHz, 50–1,000 MHz and 800–960 MHz. All units have supply voltages of 15 or 28 VDC and output power ranging from 400 mW to 1.3 W. Prices start at $52.
Motorola Semiconductor
INFO/CARD #173

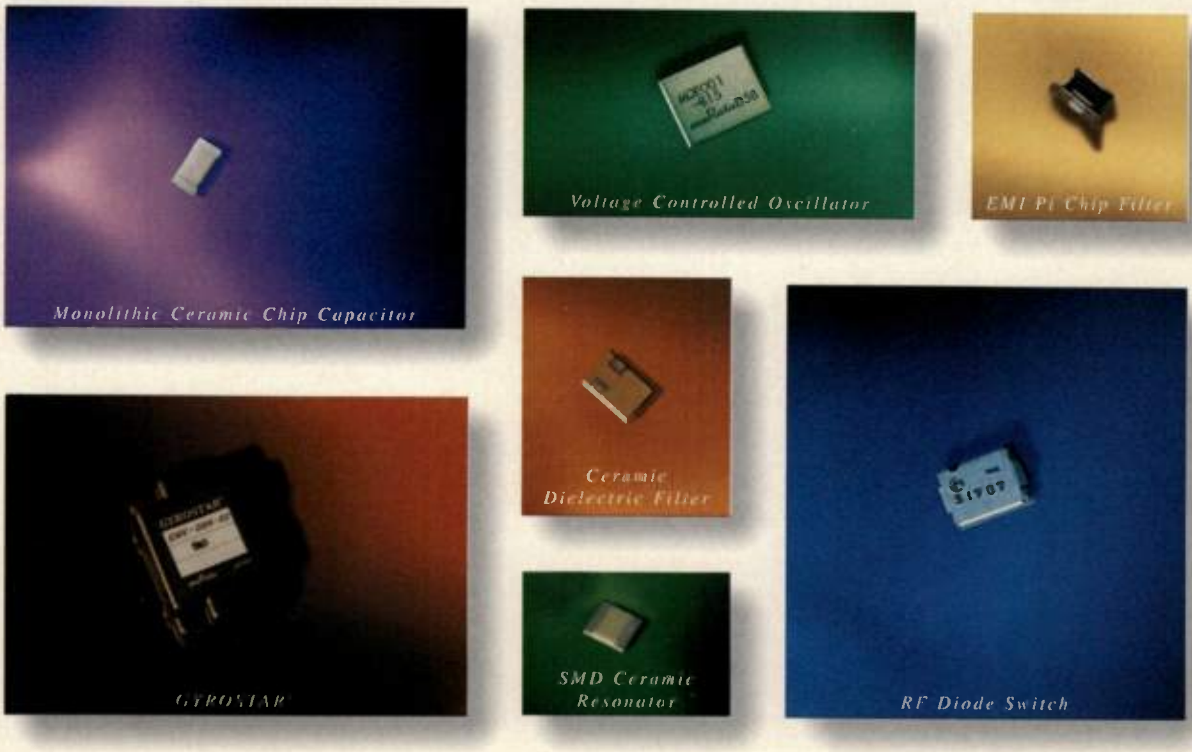
SIGNAL SOURCES

Voltage-controlled oscillator

Made by Mini-Circuits, model ZOS-1025 is a voltage-controlled oscillator with 685-to-1025 MHz linear tuning, low phase noise of –92 Bc/Hz at 10 kHz offset and power output of 8 dBm. The dual output provides well-isolated ports, and the frequency settings are virtually independent of the load. The wide range of applications include frequency monitoring, PLL circuitry, laboratory use and a frequency source for burn-in, power test and other production testing. The ZOS-1025 is priced at $119.95 in quantities of 1–9.
Mini-Circuits
INFO/CARD #174

Miniature TCXO

Model XO3062 has been added to the PTI product line for applications including portable communications systems, GPS receivers and other applications that require a small unit with low power consumption. XO3062 is a



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Our newest handbook has amplifier design information that every RF engineer needs in his or her reference library. Design principles include noise, gain and matching fundamentals, along with advanced topics like broadband matching, component modeling and feedback. All this comes together in practical design examples for front-end, IF, instrumentation and medical applications.
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high-performance, low-profile TCXO in a 1.05" × 0.69" × 0.22" surface-mount package. Models are available to cover frequencies between 16 and 75 MHz. The TCXO features ±0.75 ppm stability performance over –30°C to 70°C. The unit provides a sinewave output, and it is externally tuned.
PTI
INFO/CARD #175

High-density OCXO

Bliley offers a 10 MHz, oven-controlled crystal oscillator. Model N26S features a 7.0 dBm output into 50Ω at 1 to 10 MHz. Temperature stability is ± 0.05 ppm maximum over 0°C to 50°C. It measures 2" × 2" × 1".
Bliley Electric
INFO/CARD #176

Synthesizer

The Vari-L PLL-400-2450 synthesizer spans 2,400–2,500 MHz in 1 MHz steps. The unit requires a 5 V supply, a 3-wire serial interface and a reference oscillator input. The synthesizer is housed in a 0.6" × 0.6" × 0.138" surface-mount LCC package.
Vari-L
INFO/CARD #177

EMC

ESD, overvoltage protection array

Model SP723, an ESD-overvoltage protection array of 24 transistors in a silicon-controlled rectifier (SCR) configuration, replaces 12 discrete diodes to protect as many as six lines from 15 kV discharges. Made by Harris, the SP723 offers full suppression to Level 4 of the International Electromagnetic Capability 1000 Standard, IEC 1000-4-2. Switching in less than 6 ns, the SP723 can handle 7 A of peak current and a single 100 μs peak transient pulse of ± 4 A. The array is priced at $1.10 in quantities of 1,000.
Harris Semiconductor
INFO/CARD #178

SMT EMI filters

The 0.315" × 0.083" 4700 SMT EMI π-filter comes in a square, sectional body that saves apces and allows easy

installation with common soldering processes. The Tusonix filters are available in capacitance from 100 to 5,000 pF, and they operate at voltages as high as 100 VDC and temperatures as high as 125°C.
Tusonix
INFO/CARD #179

Compact, portable wrist-strap monitor

An important static control safeguard that monitors both the worker and the worker's wrist strap is available from 3M. The 3M 721 wrist strap monitor immediately alerts the individual when a wrist strap is malfunctioning or is operating improperly. The 721 monitor continuously emits a signal that is returned through a wrist strap and cord that contain two separate sets of independent conductors. The wrist strap monitor also checks the ground wire connection. The monitor is a compact, portable, battery-powered unit with a visual and audible alarm. The unit has a one-year battery.
3M Electrical Specialties Division
INFO/CARD #180

EMC test site characterization

A multipurpose EMC tool from Emco enables users to generate consistently repeatable electric and magnetic fields for characterizing EMC test sites. The model 4630 RefRad Reference Radiator consists of a portable, battery powered comb generator with removable transmit antennas, a remote control unit with GPIB interface, and Windows-based CalStan software for automatic data acquisition and documentation. The unit's comb generator provides manually or remotely selectable harmonic frequency spacing of 10 kHz, 1 MHz and 5 MHz over a range of 10 kHz to 1 GHz.
EMC Test Systems
INFO/CARD #181

RF software

Planning, floor layout for indoor wireless system

SMT Plus 1.0 from the Mobile and Portable Radio Research Group (MPRG) of Virginia Tech's Bradley Department of Electrical Engineering is an indoor planning tool that allows the user to design and model indoor wireless systems.

SMT Plus displays signal strength and interference contours on building blueprints for arbitrary base station placements for any indoor wireless system. Features include all primary wireless standards; plus the means to implement any number of custom standards; a capability for an unlimited number and type of base stations and interference sources inside or outside one or more buildings; simultaneous modeling of single-floor, multifloor and building penetration; rapid statistical coverage prediction methods valid at all frequency ranges and for all standards; and point-and-click access to all commands.

MPRG's interactive SiteBuilder 1.0 toolkit eliminates the difficulty of acquiring and establishing the site-specific databases used by SMT Plus by assisting in the creation of formatted building floor plans. Sitebuilder can reformat an existing SMT Plus building floor plan, format an existing building floor plan given to each floor of the building as a separate CAD drawing; format an existing building floor plan given the entire building; use a scanned image or blueprint of a building to create a formatted floor plan; or draw a formatted building floor plan from scratch. Both SMT Plus and Sitebuilder 1.0 are available for DOS, Windows 3.X, Windows NT and a variety of UNIX platforms.
Virginia Polytechnic Institute
INFO/CARD #182

Extraction modules in modeling software

Hewlett-Packard Integrated Circuit Characterization and Analysis Program (IC-CAP) Version 4.5 includes extraction modules for two key metal-oxide semiconductor, field-effect transistor (MOSFET) models. The new extraction module for the Philips MOS Model 9 is now available, and the existing extraction module for the University of California at Berkeley Short-Channel Insulated Gate FET Model (BSIM) has been updated to Version 3. IC-CAP 4.5 also includes new instrument control capabilities. BSIM3 Extraction Module now has a single equation for current and voltage through all operating regions of the device.

IC-CAP 4.5 also includes new instrument control capabilities. An instrument driver for the HP 54720 Modular Real-Time Oscilloscope has been added to the Time Domain Driver module, as well as drivers to support the HP 85124A Pulsed Modeling System.
Hewlett-Packard
INFO/CARD #183

Field simulation linked with electronic design

Ansoft's Maxwell SI Spicelink Version 3.0, a software package for skin effect and dispersion analysis, bridges the gap between electromagnetic simulation and electronic design. The software combines the accuracy of electromagnetic field solutions with the flexibility of SPICE. The software addresses signal integrity and crosstalk, parasitic parameters, power and ground plane placement and electromigration. It has capabilities for electromagnetic analysis of interconnect structures, including 3D modeling of package and die interconnects, connectors, bus structures and semiconductor metallization. Electromagnetic solvers generate circuit models of physical interconnects and then link these models to conventional SPICE circuit simulation. The software combines the modeling accuracy of the finite element method (FEM) and the efficiency of transient simulation with reduced-order models. Interfaces to Unigraphics, ProEngineer and other packages are provided through stereolithography files.
Ansoft SI Products
INFO/CARD #184

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Design engineers can locate EE information for free with Electronics Search FAQ Version 4.0 on the World Wide Web. FAQ is Internet jargon for "frequently asked questions," and the downloadable software guides users to libraries, bookstores, other engineering web sites, integrated circuit-specific resources, news-and-information sites and publications important for design engineers. Search FAQ interfaces to the search engine EE Hunter.
Eg^3 Communications
INFO/CARD #185

RF literature

Catalog selector guide details crystal oscillator

Champion Technologies recently released a new product catalog and selection guide for crystal oscillators. The catalog includes technical information and specifications on the company's line of VCXOs, TCXOs, VCTCXOs, data clocks and related products. Included is a selection guide and technical information such as a definition-of- terms section, noise definitions and product definitions.
Champion Technologies
INFO/CARD #186

Product selector, device data book updated

Motorola has released two new pieces of technical literature for RF products. Included are the latest products of Motorola Phoenix, Toulouse (France) and Hong Kong. The catalogs include power FETs; power bipolars; small- signal, monolithic ICs; and power amplifiers. Application literature and case dimensions are included.
Motorola
INFO/CARD #187

Brochure lists wireless comm products

Philips Semiconductors has published a new brochure of wireless communications products. The brochure integrates Philips' wireless product portfolio into a single document. Included in the brochure are product specifications and block diagrams. Sections are arranged by application.
Philips Semiconductors
INFO/CARD #188

Components report focuses on European market

A report detailing the European microwave components industry has been released by Miller Freeman. It contains detailed information about the technology and market trends expected to affect the industry during the next five years. The report defines the technology and discusses aspects of the European market, its relation to the global economy and the market's strengths and weaknesses. It analyzes trends and presents data in tabular and chart formats, allowing for easy comparison of products, markets, applications and locality.
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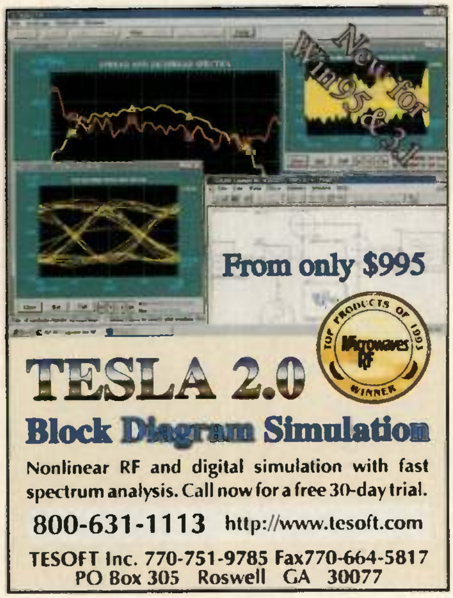

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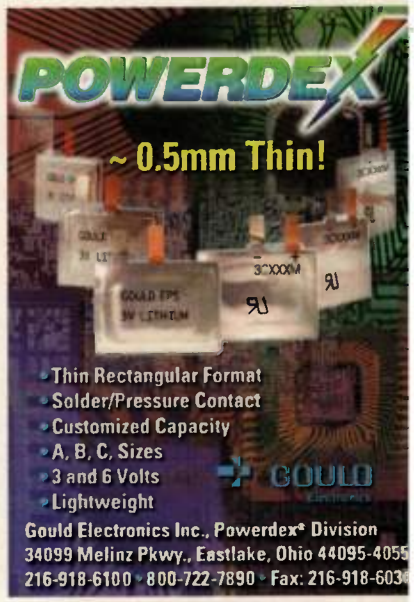

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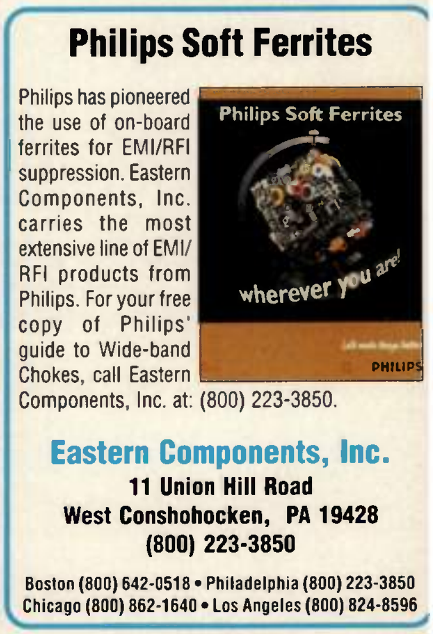

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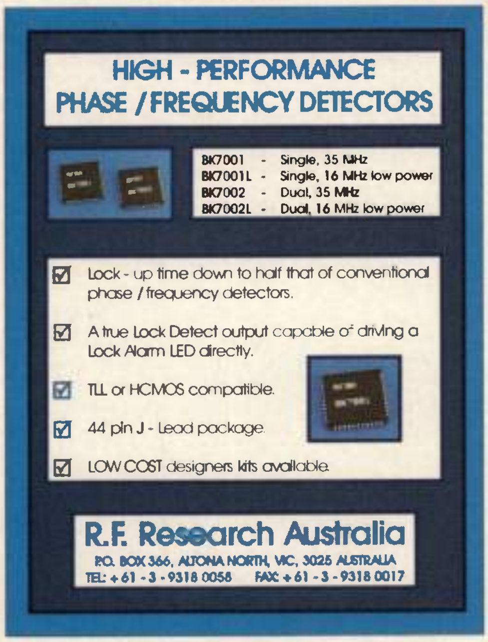

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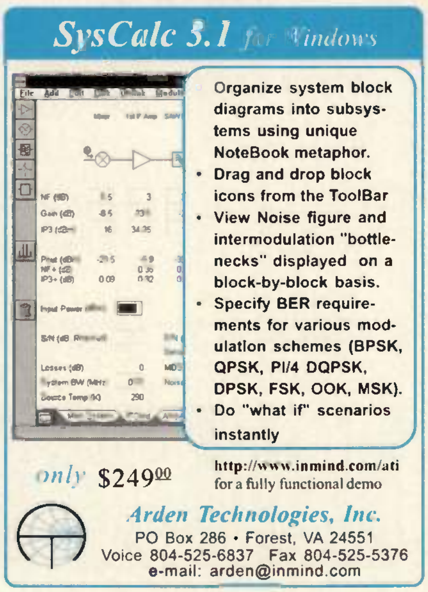

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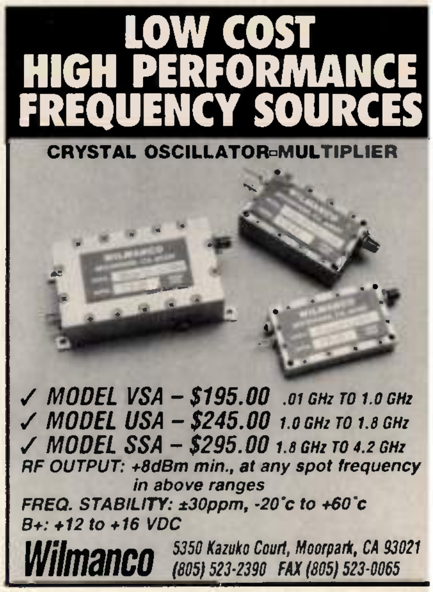

INFO/CARD 108

RF marketplace

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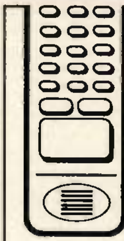

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Regional Field Sales: Aggressive individuals to create and serve new accounts. Positions are located throughout the U.S.A. An engineer who wants to enter sales world is acceptable. Base salary, commission and car. BSEE.

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Sr. Project Antenna Design: Lead the conception, design and development of a wide variety of antennas and antenna systems, including both reflector and array systems using microstrip, stripline and waveguide technologies. BS/MS with 5 years experience.

RF Design Manager: Lead a team of RF engineers from initial design and implementation through product integration and testing into high volume production. 8+ years of RF design with emphasis on low cost radio design. BS/MS.

Sr. MMIC Design: Design highly integrated GaAs MMICs for advanced cellular products. Circuits to be designed include: power amplifiers, LNAs, mixers, IF amplifiers, buffer amplifiers. RF frequencies are 900 and 1800 MHz.

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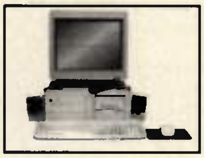

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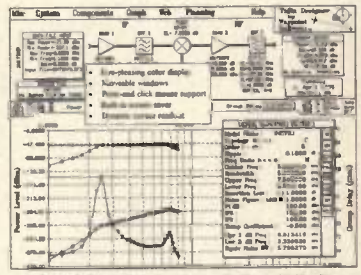

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JMS-1LH	+10	2-500	DC-500	5.75	55	45	8.45
JMS-1MH	+13	2-500	DC-500	5.75	60	45	9.45
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JMS-2	+7	20-1000	DC-1000	7.0	50	47	7.45
JMS-2LH	+10	20-1000	DC-1000	6.5	48	35	9.45
JMS-2MH	+13	20-1000	DC-1000	7.0	50	47	10.45
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